Final Report

for

STUDY PROGRAM ON (30 - 100 GHz) ELECTRONICALLY STEERABLE ANTENNA SYSTEMS

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ABSTRACT

Millimeter wavelength systems have become increasingly attractive for use in spacecraft communication systems. The purpose of this program was to explore the possible techniques available for an electronically steerable antenna system and build a feasibility model of the system or systems which were most promising.

The first part of the program was concerned with a detailed examination of several means of achieving the goal based on lower frequency microwave methods and higher frequency optical techniques. Of the many examined, two systems were chosen as having substantial merit and were selected for further study. These two were a conventional waveguide method of feeding an egg crate array and a quasi-optical one for feeding a similar array. Feasibility models of the two systems were fabricated and tested. It was found that the former was in excellent agreement with the predicted performance at 35 GHz but would be limited at 94 GHz because of increased waveguide losses. Although the latter was not completely developed at the close of the program, experimental results indicated that it would be less frequency sensitive and have lower losses at the higher frequencies. In view of this, it was strongly recommended that development of the quasi-optical feed system be continued. The mathematical expressions developed in this report facilitate the redesign of the system for applications other than the one considered in the present contract.

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1. INTRODUCTION

Millimeter wavelength systems are becoming increasingly attractive for use in spacecraft communication systems. Their smaller physical size and weight combined with their inherent larger bandwidth capabilities make them an ideal solution to many problems encountered in the space program. However, the state-of-the-art of components in this frequency range has hindered its exploitation.

The purpose of this program then, was to explore the possible techniques presently available for an electronically steerable antenna system and build a feasibility model of the system or systems which were most promising. The goal for this program was an antenna centered at 35 GHz with a 1 GHz bandwidth. The gain of this antenna was to be 30 dB above an isotropic radiator and it was to have a $\pm 12^{\circ}$ steering capability.

An obvious beginning for such an investigation was a survey of the existing techniques for lower frequency electronic scanning arrays in the light of their potential use at millimeter wavelengths.

The evaluation was conducted on the basis of four separate functions of a steerable antenna system:

1) The electronic means to steer the beam.
2) The distribution system which feeds energy to the array.
3) The structure of the array of phasing and radiating elements which satisfy the requirements of coverage, beamwidth, environment, etc.
4) The control of beam pointing either locally or remotely for proper coverage and tracking.

Of the many techniques that come under each heading, certain ones are more readily adaptable to millimeter wavelength operation. Combinations of the methods from each of the four function categories provides a total electronically steerable antenna system. Six combinations were selected for further study and are described in a later section of this report.

Many difficulties are encountered when scaling microwave techniques to the millimeter range (e.g. critical dimensional tolerances, losses inherent in transmission line etc.). However, recent advances in solid state component technology at millimeter frequencies together with the fact that a waveguide array can be constructed in a single process (i. e. electroforming techniques), provide definite potentials for scaling the microwave techniques into the millimeter bands. Since optical techniques can also be used in this range, an ample number of possibilities for investigation exist.

2. SYSTEMS STUDIED

2.1 Electronic Steering Techniques

Several low frequency techniques for steering were rejected immediately, i.e. Luneburg lens systems which have material problems yet to be solved, and frequency scanning methods which are restricted by the application being considered, to mention two. Others held considerably more promise and were studied farther.

2.1.1 Scaling Low Frequency Techniques

The formation and steering of a beam at microwave frequencies with antenna arrays has been accomplished by controlling certain parameters of the aperture. The shape and direction of a beam depends, for the most part, on the size of the array, the number and spacing of the elements, and the excitation function. This last term describes the relative amplitude and phase of the signal at each of the radiating elements. The beam steering capability is derived from the phase function whereby the angular relationship between the transmitted wave front and the plane of the array can be changed to point the beam in a desired direction. When variable phase control is introduced at each element, phase shifter array steering is achieved and the array can be fed by a distribution network having a single input port. If the proper wave-front phasing is generated in a constant network (e.g. Butler matrix), then the array is energized by multiple feeds and each feed port corresponds to a beam position. Finally, it is possible to vary the phase effectively by changing the electrical length of feed between array elements with a sweeping frequency. However, the present application precludes this method. With these three basic techniques as a base, then, several different configurations are possible.

2.1.1.1 Phase Shifter Methods

The most popular electronically steerable antenna technique is the phase shifter array. This technique has been extensively utilized in the microwave region but in the millimeter bands it has been difficult to implement. A limited steering of planar arrays has been achieved by inserting phase shifters in the feed arm between the rows of linear slot arrays. This allows steering in one dimension only. Further growth of this scheme to scanning in a solid angle requires that a phase shifter be placed at each slot element rather than just between each row. It therefore becomes necessary either to enlarge the spacing between the slots for a series feed, or feed the slots with a parallel device such as a corporate structure positioned behind the array,[1] or switch to a non-closed waveguide distribution system.

When the aperture takes on the form of an array of series-fed slots or parallel-fed open-ended waveguide, the aperture will begin to resemble a discrete lens which is sectionalized in two dimensions. Such a structure is sometimes called an "egg-crate" array due to its appearance. In some respects the modular construction characteristic of the egg-crate offers some advantages not found in other configurations. For example, an egg-crate structure has a natural compactness and duplication of sub-array units to meet the needs of a millimeter antenna system. It is also possible to do away with the conventional types of feed structure for an egg-crate and use quasi-optical designs which could prove to be more convenient for millimeter wavelengths.

The egg-crate array is well-known and has been used for fixed lenses with success.[2] A ten-element egg-crate lens structure has been built and tested at Beckman - ADTEC as a

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1 A linear array of this type has been built for 50 GHz operation by Republic Aviation Corporation on Air Force Contract AF 33(657)-8855.

2 Brown, J., Microwave Lenses, Methuen, page 63, 1953.

harmonic generator array.[3] Diodes in each of the channels of the structure produced a 2:1 frequency multiplication of the signal traversing the array. Further discussion of this technique appears in the section on Up-Conversion Array Techniques. The experimental results of a 104 element egg-crate reflector array at 7080 MHz also have been reported by Malech, Berry, and Kennedy.[4] Variable shorts were used as the phasing elements in what was called a "reflectarray" antenna.

Both the transmission type egg-crate structure and the reflection type may be electronically steered by placing phase shifters in each compartment of the egg-crate. There are three types of electronically controlled phase shifters that have possibility for performing satisfactorily at millimeter wavelengths - the continuous ferrite, the latching ferrite and the diode. The latter two devices normally provide digital phase shifting. For the most part all three devices are in an early stage of development at the upper end of the millimeter band but are commercially available for frequencies close to 35 GHz. From the standpoint of adaptability to the egg-crate structure and convenience of operation on board a satellite, the ferrite latching device offers the most promise. These devices use ferrite elements that can be switched from one saturation state to another by a video pulse and remain in the new state between pulses. The result is the discrete shift of phase of the millimeter wave energy passing through the guiding structure. A set of such ferrite elements in a single guide constitutes

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[3] King, D.D., Sobel, F., Dozier, J.W., "Harmonic Generation by An Array", IEEE Trans. on Microwave Theory and Techniques, Vol. MTT-12, No. 2; March, 1964.

[4] Malech, R.G., Berry, D.G., Kennedy, W.A., "The Reflectarray Antenna System", 12th Annual Symposium USAF Antenna Research and Development Program, University of Illinois; October 16-19, 1962.

a digital phase shifter which can be controlled directly by digital circuits. As a result there is a natural interface with a digital computer controlling the beam pointing and the drive power required to change the phase is much less than that required by the analog ferrite phase shifter. The only parameter of the latching device which needs to be specified is the quantization of the bits necessary to give satisfactory beam pointing.

2.1.1.2 Multiple Beam Methods

Many of the beam forming systems steer the beam by either mechanically positioning the entire aperture or by switching between multiple feeds to the beam forming device, thereby generating a number of possible beam positions. The electronically steerable ones include the various lenses and the closed circuit matrices such as the Butler network. In these cases electronic scanning is provided by switching between off-axis feeds. From the point of view of an external observer the behavior of the scanning beam from an array employing a beam forming device is no different from the phase shifter array with digital phasing.

The beam forming devices are the common denominator in this category. A switching network is required in all cases. The devices perform the same function (modification of phase), and lenses, being continuous, offer the advantage of absolute tolerance relaxation. Lens techniques are based on the principles of geometrical optics. Open wave illumination is generally used to feed such an aperture. Further discussion of lenses appear in the Section on Quasi-Optical Designs.

In the closed-waveguide category of the multiple beam techniques, there are also several methods of forming a beam. Among the most popular are the Butler beam forming network,[5] the multiple beam interval scanner (MUBIS)[6] and the Maxson array.[7] In each of these systems the beam forming is implemented at the microwave frequency so that the networks are reciprocal.

The Butler beam forming network is a matrix composed of hybrid junctions and fixed phase shifters arranged in a fashion which produces a number of beam positions equal to the number of elements in the array. Millimeter wavelength hybrids and fixed phase shifters exhibiting extremely low loss characteristics are necessary since so many are involved. The loss just due to the complexity of the switching matrix is a sufficient drawback for this technique at millimeter wavelengths to warrant disregarding its possibility.

The MUBIS antenna system uses a parallel plate lens to form the multiple beams. Feeds are placed at one end of the plate and the corresponding outputs at the other end are connected to the array. The parallel plate region of the MUBIS system performs a similar function of beam separation as in the Butler network. Although the transitions into and out of the parallel plate region introduce some loss and reflection, the major drawback is, again, the complexity of the switching matrix necessary to gain access to the feeds ports for different beam positions.

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[5] Butler, J., Lowe, R., "Beam-Forming Matrix Simplifies Design of Electronically Scanned Antennas", Electronic Design, Vol. 9 pp. 170-173; April 12, 1961.

[6] Sylvania Final Report, "Multiple Beam Interval Scanner", Contract AF 19(604)-7385; March 23, 1964.

[7] Transactions of 5th DOD EW Symposium, University of Michigan, pp. 125-129.

The Maxson system is a traveling wave type configuration which uses a directional coupler network to produce multiple beams. It is not an orthogonal network in the sense of the Butler network, i.e. it requires dissipative terminations to absorb unwanted components of the signal and is thereby slightly inefficient. Here again, past work has been done at much lower frequencies. The size and bulk typifying the structures that have been developed (phasing was obtained by the use of lengths of transmission line) is a reasonably good indication of what would be expected at millimeter frequencies. The complexity is generated by the use of N^2 directional couplers which all have a different degree of coupling; a factor which is significant enough to disallow this technique from being considered.

In general, the closed waveguide multiple beam approach is less appealing than the other techniques that have been considered. The quantity of components alone in most of the methods becomes a major obstacle in the design of a small size, light weight antenna package. Loss problems are also significant and for the most part are at least equal to those losses associated with the phase shifter techniques.

2.1.1.3 Up-Conversion Methods

Two possible ways of avoiding the increased attenuation at the high millimeter wavelength frequencies make use of solid state diodes as multipliers and modulators. The first would consist of an array of multipliers for which the distribution and phase shifting is done at a subharmonic of the radiated frequency. The second would use an array of single sideband modulators. The IF phasing signal, which is of no consequence to the efficiency of the overall system, is injected into the array of balanced modulators from which the upper sideband is tapped and radiated from the array steered in a direction determined by the phasing introduced at the IF. This system is very appealing but, due to the lack of an efficient single sideband modulator at 94 GHz, it does not appear about to become a reality.

Both of the above schemes are not totally new approaches to beam steering, but are instead methods designed to aid in the implementation of the phase shifter array techniques at millimeter wavelengths.

2.1.1.4 Retrodirective Systems

Since Van Atta first derived the retrodirective principle for an array of antenna elements,[8] many different schemes based on the self phasing principle have been presented. The variations have included first techniques in which signals received across the array are transmitted after being processed individually (that is, individual signals are not combined with each other prior to retransmission). Besides the Van Atta array this category includes the feedback phase-shifter; double heterodyne and single heterodyne techniques. Other schemes have made use of phasing matrices or beam forming networks to receive the signal in the same or selected directions and to retransmit in that same direction. A third category makes use of those techniques which adaptively optimize the performance of a receiver array to then be used as the transmitting antenna. Of the techniques investigated, the single heterodyne using the lower sideband output of an up-converter seems to be the best selection for a millimeter wavelength application.

In the single heterodyne scheme,[9] the converter is driven with a frequency that is approximately twice the

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[8] Van Atta, L.C., Electromagnetic Reflector, U. S. Patent No. 2,908,002.

[9] Rutz-Philipp, E.M., "Spherical Retrodirective Arrays", IEEE Trans. AP-12, pp. 187-194; March, 1964.

incoming signal. The lower sideband then is the same as the incoming signal displaced by a slight frequency offset. By using branching filters and circulators to isolate the input and output signals, the system tends to become more stable than it was in the approach referenced above. The converter may be a lower sideband (inverting) up-converter producing a controlled amount of negative resistance gain. By using the lower sideband, the phase of the incoming signal is conjugated and appears as the negative phase on the lower sideband.

2.1.2 Quasi-Optical Designs

By definition there are some techniques that lie more closely akin to the principles of geometric optics. Because the relationship falls in this span of the electromagnetic spectrum these techniques are fundamentally quasi-optical designs. Generally they are multiple beam forming systems with unbounded illumination. The paraboloidal reflector with a single or multiple feed is the most commonly used approach in the geometrical-optic family. Very little need be said further about this method since its characteristics are well known. The associated problems of aperture blockage and limited electronic scan capability are magnified still further in a millimeter wavelength application. Quasi-optical methods that have more potential at the millimeter wavelength frequencies are the lenses including the Luneburg, constant-dielectric, ferromagnetic and Fresnel zone-plate types of lenses.

2.1.2.1 Dielectric Lenses

Various lens configurations have been used to provide electrical beam steering. The well known Luneburg lens principle of optics has been applied quite frequently to frequencies below X-band. The Luneburg lens is a sphere of dielectric material in which the dielectric constant increases continuously from that of air at the periphery to twice this value at the center. The rate of increase is such that a plane wave striking the lens from any one direction is focused at a point on the opposite side of the sphere. A feed placed at this point will transmit the energy which passes through the Luneburg lens and travels in the selected direction. A mosaic of feeds about this point will produce beams in adjacent solid angle directions. If the adjacent feeds are sufficiently close, their corresponding beams will overlap. In this way a nearly continuous coverage angle can be made available to the steerable array system.

The Luneburg lens has drawbacks which at the present appear to be insurmountable. A continuously changing dielectric constant must be approximated by a series of concentric shells with stepped dielectric constants. The present fabrication techniques and materials used by manufacturers such as Emerson and Cumming and Armstrong Cork lead to quite lossy devices at frequencies above X-band. Another factor against the Luneburg lens is the fact that practical fabrication of the outside shells is not feasible if they are less than 1/8 inch thick (which happens to be one wavelength at 94 GHz). It is this extreme difficulty of controlling the dielectric constant and dimensions in the small steps that are necessary at millimeter wavelengths that place the upper boundary of this technique at frequencies in the neighborhood of 18 GHz at the present time.

The fabrication problems of the Luneburg lens are solved to some extent by the constant-dielectric lens technique. However, the weight and loss problems are equal if not greater than the Luneburg lenses. The constant-K lens operates in a manner by which the rays focus, through refraction, at controllable points lying from within the surface of the lens to the perimeter.

The position of the focal point is dependent on the dielectric constant of the lens material. These lenses are simpler to construct than the Luneburg but have not been put into use above X-band. It appears that here again the loss problem is a major obstacle to be overcome for applications at millimeter wavelengths.

The problems of weight and loss are not encountered with thin lenses to the extent that they are with the spherical lenses. However, the errors due to aberration prevent the scanning angles from becoming too large. The monochromatic aberration known as coma is the inability of a thin lens to focus at off-axis points. Thin lenses depend on the refraction at the boundary between different dielectric media to perform the focusing function. A technique which utilizes the interference between annular grooves on a dielectric plate to perform the focusing is known as a Fresnel zone plate lens.[10]

The simplest and historically the earliest form of the zone plate consisted of a plane array of alternately opaque and transparent concentric circular rings. A screen in this form acts upon a normally incident plane wave transforming it into a converging wave which is approximately spherical, thus concentrating the radiation in the small region about a point on the axis. A more advanced form of zone plate operates on the phase of the energy impinging on the opaque regions such that upon reaching the other side of the dielectric it adds constructively with the energy transmitted through the original non-opaque regions.

A phase correcting Fresnel zone plate with the properly designed feed horn at its focal point constitutes

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[10] Harvey, A.F., "Optical Techniques at Microwave Frequencies", Proc. of the IEE, Vol. 106, Part B, p. 141; March, 1959.

an antenna. The addition of off-axis feeds in a mosaic would provide a multiple beam antenna in which the pointing direction of each beam depends only on the location of the feed with respect to the center of the lens. The displacement of the feeds from the focal point is limited to a few beamwidths if serious beam deterioration is to be avoided. For extremely large scan angles it would be necessary to move the entire antenna structure.

The zone plate is made much thinner than the equivalent refracting lens which results in less transmission loss and a decided advantage in weight. The zone plate lens would weigh one-half to one-third of that of the conventional lens using the same dielectric material and of course considerably less than the solid Luneburg and constant dielectric lenses. The transmission loss for a quarter-period lens is less than 0.5 dB over a large part of the millimeter frequency range.

All of the techniques presented in this section have been discrete systems in which the beam steering is accomplished by the use of multiple off-axis feeds to form multiple beams. As a result, this method of steering has limited scan angle capabilities because of the spherical aberrations that are encountered through the use of off-axis feeds. The next section presents a scheme which makes use of the bulk effect of ferromagnetic material to direct the beam over the desired solid coverage angle.

2.1.2.2 Ferromagnetic Lenses

Ferromagnetic materials which exhibit a phase tapering quality when placed within a magnetic field can be used as a lens when a feed horn emitting a plane wave illuminates the ferrite material. The beam is shifted in proportion to the magnitude of the field inserted from the lens edges. Models of the technique [11] have demon-

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[11] Johnson, R.E., Schiller, T.R., Weiss, P.F., "Ferromagnetic Beam Steering at Millimeter Wavelengths", IEEE Int. Symposium PTGAP pp. 71-77; September, 1965.

strated a capability for steering a beam in one dimension by $\pm 20^{o}$ at 71 GHz. A total applied field variation equivalent to 144 ampere turns was required to steer the beam between the 20^{o} limits. For small apertures the magnetic field requirement is sufficiently small to lend some appeal to this technique for the satellite application.

One inherent property of this scheme, restricts the use to circular polarization. Another challenge to optimum performance is the hysteresis effect of the ferromagnetic material. A reversal of the magnetic field does not return the beam to the same position it occupied before the reversal. Another problem at the present time is that only small pieces of ferrite of the type needed are available. Maximum dimensions of commercially available ferrite are in the neighborhood of 1/2 by 1 by 6 inches. A 5 inch aperture therefore must be made up from a number of small pieces cemented together. As an alternative to this complication the investigators directed their thinking towards obtaining the high gain of a wide aperture by forming an array of smaller ferrite scanners. The development along this line appeared to be following a growth pattern leading toward the conventional phase shifter method in which control of the beam direction is accomplished at each subaperture of the antenna.[12] In view of the results that were being obtained with the ferromagnetic lens this approach was not pursued during the present study.

2.1.3 Systems Selected For Further Study

The previous discussion has illustrated that although there are an abundance of techniques which for one reason or another are impractical there are still several which demand further study.

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[12] Schiller, T.R. et. al., "Electrically Steerable Array", Sylvania Tech. Report; November, 1965, AD 624 791.

Most promising of all the schemes discussed is the phase shifter array approach. This would entail having a phase shifter at every element in the array thus making possible solid angle steering. The ability of each phase shifter to change its degree of phase shift by electrical control is a necessary parameter. Diode phase shifters which switch in fixed line lengths are very lossy at the frequencies in question and require the use of circulators in a transmission type array (as opposed to a reflectarray). Ferrite phase shifters, although also lossy, appear to be more practical since they lead to a lighter; less complex system. In addition, by judiciously choosing the number of bits in each phase shifter any degree of pointing accuracy can be achieved together with a nearly continuous scan. These phase shifters are still largely in the development stage at the present time but are feasible even at 94 GHz.

The phase shifter array is superior to all multiple beam methods due to its more continuous scanning capability and its freedom from the constraining feed switching situation inherent in the latter method. These advantages make it very appealing and worthy of additional investigation.

Closely akin to the phase shifter method, the up-conversion methods hold a considerable amount of merit. Solid state multipliers at these frequencies are within easy reach, though single sideband balanced modulators are still experimental and need much more development. Therefore the method that injects its phase control at the fundamental frequency is the most practical first approach to investigating this technique.

TABLE I

SUMMARY OF SECTION 2

STEERING METHOD	REMARKS
Phase Shifter Array	Adequate and fairly easy to implement but limited by the availability of the phase shifters. However, the most practical system considered.
Butler Beam Forming Network	Losses and complexity too great at millimeter wavelengths.
Multiple Beam Interval Scanner	Complexity of switching matrix to off axis feeds for different beam positions is sufficient to disregard the system.
Maxson Array	The required use of N^2 directional couplers each with a different degree of coupling is unnecessarily cumbersome.
Harmonic Array	Promising. Further investigation recommended.
Balanced Modulator	Promising, but lack of an efficient single sideband modulator at the frequencies in question precludes its use.
Retrodirective Array	Unnecessarily complex for the application being considered.
Dielectric Lens System	Fabrication and weight problems, material problems, and requires switching to off axis feeds.
Ferromagnetic Lens System	Rejected because of restriction to use of circular polarization; availability of ferrite material; large magnetic field necessary.

3. DISTRIBUTION SYSTEMS

Many types of low loss millimeter wave distribution systems were examined, and complete descriptions of the two most promising are presented in this section. These are a conventional waveguide technique and a quasi-optical illumination technique.

3.1 Waveguide Networks

The ordinary method of distributing energy at microwave frequencies is hollow metal waveguide. The basic dominant mode rectangular waveguide used in these distribution networks may be reduced in height for structural convenience or to provide compact delay sections. Unfortunately, at millimeter wavelengths, the attenuation is appreciable and the dimensional tolerances required to maintain the proper phase relationships are tight. As frequency is increased and the transverse waveguide dimensions are correspondingly decreased, the theoretical attenuation varies as $f^{3/2}$. If the line lengths are normalized to wavelength, the total insertion loss theoretically increases as $f^{1/2}$. In practice, surface roughness effects in commercially available waveguide cause the insertion loss to increase at an even faster rate. Recent advances in electroforming techniques have helped to reduce the roughness problem. Hence, if the line lengths are small, it is still feasible to use this type of distribution system at least at 35 GHz where the loss encountered in dominant mode waveguide is about 0.2 dB per foot. At 94 GHz the loss increases to about .75 dB per foot. Together with the additional loss introduced by power dividers such as tee-junctions the losses start to become prohibitive at the higher frequency. However, due to the simplicity and straightforward design of this system it was worth the effort to examine it further.

3.2 Quasi-Optical Technique

Another technique which, though simple, looked promising is that which employs quasi-optical illumination as the feed system. In this arrangement a large pyramidal horn is used to illuminate an egg crate structure. Phase shifters located in each segment of the egg crate are then used to modify the phase distribution of the transmitted signal and enable steering in a solid angle. The advantages of this technique are many. Waveguide wall losses are greatly reduced by feeding with a single horn and, intuitively speaking, the system would have an inherently greater bandwidth. In addition, the weight of the quasi-optical system would be less than the weight of the conventional waveguide network.

The system could be implemented in two different ways. In the first method, the walls of the large illuminating horn would not extend out to the egg crate and the space separating the two would be surrounded by absorbing material. This method would be slightly lossy since energy would be lost due to spill over, i.e. side lobes of the illuminating horn which are not captured by the egg crate structure. The second method would correct this by extending the walls of the illuminating horn out to the egg crate thus completely enclosing the energy. Although this would reduce losses it would also trap unwanted reflected energy and higher order modes inside the large horn. Nevertheless, whichever system would be better, the quasi-optical feed technique had sufficient merit to warrant further investigation.

4. APERTURE CONFIGURATION

A basic consideration for a discrete millimeter wavelength antenna system is how well the technical difficulties involved in constructing an array of elements can be overcome within dimensional limits that are relatively minute. Half-wavelengths on the order of 1/16 inch at 94 GHz, for example, present a rather complex problem for realizing suitable radiating elements and element spacing. Fully steerable antenna arrays with gains of 30 dB require the use of hundreds of elements. With the limited spacing and tolerance, the problem becomes even more severe.

4.1 Array Techniques

It is generally known that greater versatility can be achieved with an antenna array than with a single element. Electronic steering of the main beam, difficult if not impossible with a single element, is easily accomplished by an array by varying the relative phase at each element. Since one of the objectives of the present program is electronic steering, an array is the immediate choice.

There are four parameters which need to be specified when designing an array. These include: size of the overall aperture; the individual elements; the spacing of these elements and the illumination function which describes the relative amplitude and phase at each element. To meet the specification of this program that the antenna have 30 dB gain above an isotropic radiator requires that the aperture be at least 15 wavelengths square. At 35 GHz this would mean an aperture size of approximately 5 in. on a side while at 94 GHz the size would be 2 in. by 2 in. The customary spacing of the elements is 0.6 wavelengths which insures that the grating lobe does not appear until extreme steering angles are attempted (i.e. angles close to 90° off broadside). This restriction can be relaxed since the present application requires only $\pm 12^{\circ}$ steering. Hence the element spacing can be increased. However, to insure maximum gain and minimum decrease in gain when steered off boresight requires that the aperture be filled even though the element spacing can be increased (i.e. the number of elements decreased).

4.2 Array Elements

From the above discussion it is seen that the collection of individual radiating elements would ideally have the ability to fill completely the aperture and be larger than .6 wavelengths square. In addition, the radiation pattern of the array element must be considered since it determines to a large extent the angle over which the beam can be steered. Large scan angles require broad radiation patterns and therefore lower gain elements. Consequently, the reduced steering requirements in this program allow the use of a smaller number of higher gain elements. One type of radiating element that can meet these needs is the pyramidal horn. Since its aperture can be varied, it can be expanded to completely fill the overall aperture no matter how many elements are needed. At this point the only two parameters which remain to be specified are the spacing of the elements, which determines the quantity of elements, and the illumination function. These entail a detailed analysis and will be treated in a later section.

5. CONTROL SYSTEMS

The prior sections have categorically considered the major functions of an array system. As stated in the Introduction, the selection of an optimum antenna system for both 35 GHz and 94 GHz consists of at least four parts. One part is the selection of the electronic means for steering the beam (Section 2). The second is the selection of the feed which joins the input terminal to the array (Section 3). The third is the structure of the array of phasing and radiating elements to satisfy the requirements of coverage, bandwidth, etc. (Section 4). The fourth is the control of beam pointing either locally or remotely for proper coverage and tracking. The latter function is dependent on the components and steering techniques selected and was not considered in detail in the early phases of the program. Consequently, it will be treated in a later section.

6. DEVELOPMENT OF SELECTED SYSTEMS

As a result of the survey of the various approaches to the problem as outlined above, several techniques were eliminated in the light of the state-of-the-art of component technology at the frequencies of interest. The following section describes the development of those systems which appeared to be the most promising for either or both frequencies (i.e. 35 GHz or 94 GHz).

6.1 Radiating Aperture

Since the primary objective of the program is the specification of the antenna itself, the first logical step was to determine its design and then use the requirements of this configuration as the design parameters for the feed system.

There are a large number of variables involved in the design of the antenna and significant trade-offs between them. These include the overall aperture size and hence beamwidth; the individual element gain and its effect on the steered aperture gain; the element spacing which determines grating lobes; the steering requirement and its interaction with the element spacing and the illumination function. These are discussed below.

6.1.1 Aperture Size and Element Spacing

One of the requirements of the antenna is that it have a gain at least 30 dB above an isotropic radiator. The gain of a completely filled array is determined by $4\pi A_{eff}/\lambda^2$. Assuming a uniformly illuminated square aperture with a 16λ side and 60% efficiency, this computes to 33 dB. The 3 dB beamwidth for such an aperture is approximately λ/D, where D is the side of the square, or about 3.5°. These figures were compatable with the goals and were used in subsequent calculations. There remains the problem of the minimum number and spacing of the elements comprising the array to satisfy the coverage requirement.

The element spacing determines the position of the grating lobes for the array. To avoid possible ambiguities while achieving coverage of the earth, it would be necessary to have the first grating lobe such that it would never point at the earth. The relation between steering angle at the antenna and elevation angle on the earth is

$$\sin \theta = \frac{r}{r + R} \sin \psi$$

where

r	=	earth radius
R	=	orbit altitude
θ	=	steering angle off boresight
ψ	=	elevation angle measured from the vertical

Since the attenuation through the atmosphere follows a secant law quite closely, the practical upper limit on ψ is something less than 90°. Keeping this in the range of $80\text{-}85^{\circ}$ gives a maximum value for θ of approximately $8^{\circ}\,40'$ for the case of synchronous orbit. This in turn implies the first grating lobe of the array should be more than $17^{\circ}\,20'$ off boresight. In effect this sets the lower limit on the number of elements to a 5 x 5 matrix.

For the initial calculations of the far field pattern a simplifying assumption was made - namely that the phase and amplitude distribution across the array were uniform. This led to simple integrals when taking the Fourier transform and allowed some hand computation. An additional assumption was that, since the earth subtends approximately $\pm 8.5^{\circ}$ from a synchronous orbit, and there can be some error in the orientation of the platform, the region of most interest in the antenna pattern is that of the $\pm 12^{\circ}$ region as measured off broadside.

The results of these computations indicated that the

5 x 5 matrix was unsatisfactory since a portion of the grating lobe moved into the region of interest when the antenna was steered 8° off broadside and for the same reason the 6 x 6 element array was just barely acceptable. Attention was then focused on the 7 x 7 element array which appeared to be acceptable under the original assumptions. However, in practice, the phase front is not uniform at the aperture of a pyramidal horn but in fact has a parabolic contour due to the nature of the propagating modes in the horn. Similarly the amplitude is not uniform across the aperture since in the E-plane of the horn it must go to zero at the walls. Finding transforms of this type of distribution then leads to integrals of the form

$$I_j = \int_{a_j\lambda}^{b_j\lambda} \begin{Bmatrix} \sin \\ \cos \end{Bmatrix} \left(\frac{\pi n}{16\lambda} z \right) \exp\left\{ i \frac{2\pi}{\lambda} \left[z \sin\theta + z \sin\phi - \frac{n^2}{256\pi\lambda} \left(z - \left(\frac{a_j + b_j}{2} \right) \lambda \right)^2 \right] \right\} dz$$

where now n is the number of elements and a_j and b_j are constants which will be bounded by $-8 \le a_j < b_j \le 8$. The cosine function is used in the integral when the element number is odd, and the sine function when it is even. The field strength for any pattern will then be given by $\sum_{j=1}^{n} C_j I_j$ where C_j is the weighting factor for the desired amplitude distribution, which for first considerations was constant. The H-plane pattern was derived from the same type of integral with the sine or cosine term replaced by a constant.

Since the sidelobes of a uniformly illuminated aperture are only down 13.2 dB in the unsteered condition and get worse with larger steering angles and since this condition could lead to ambiguities, something had to be done to reduce the sidelobe level. This lead to adjusting the final aperture parameter - the illumination function.

6.1.2 Illumination Function

The illumination function of an array completely specifies the phase and amplitude at each element. The individual phase for a broadside beam (i.e. unsteered) is uniform across the array and changes only when steering the beam. The amount of change at each element is a function of the spacing of the elements, the position of the element in the array and the steering angle desired. For a uniformly spaced 7 x 7 element array with a 16λ side the elements would be located on .771 inch centers (at 35 GHz). The amount of phase shift needed at the nth element in any row to steer the beam θ^o is given by

$$\Delta\phi_n = n\,\Delta\phi' \quad n = 0, 1, 2, \ldots 6$$

where $\Delta\phi'$ in degrees is given by

$$\Delta\phi' = 360\left(\frac{16}{7}\sin\theta\right)$$

where θ is the steering angle. The same expression is valid for the phase shift for each element in any column. The final required phase shift is then the sum of the two.

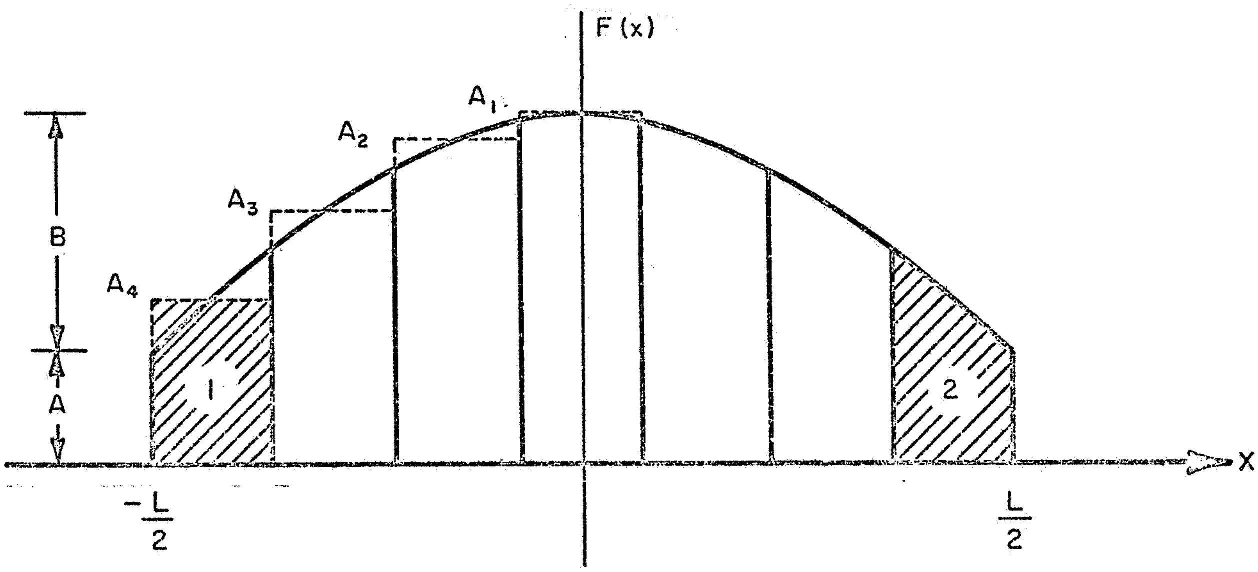
The amplitude term of the illumination function still remained to be specified. Since it is this term that has primary control of the level of side lobes, some distribution other than a uniform one was necessary to reduce their level. There are several distributions which will accomplish this and all involve tapering the distribution such that the maximum power is radiated from the center of the array and is gradually diminished out to the edges. A characteristic of a tapered illumination is that it reduces the sidelobe level at the expense of broadening the main beam. Thus there is a trade-off involved.

One distribution which reduces the sidelobe level considerably is a cosine on a pedestal. In this scheme more effective use of the full aperture can be maintained even when the beam is steered since the taper does not go to zero at the edges (which would further reduce the sidelobe level). However, the smooth cosine curve must be approximated by a stepped taper since with discrete elements a continuous distribution is impossible.

The method of dividing the power up across the aperture was to treat the problem geometrically as illustrated in Figure 1. For a 10 dB amplitude taper it can be shown that A and B are related by B = 2.16A. Then if we require that the hatched areas 1 and 2 be equal (and similarly for the remaining two pairs and the center strip), the results showed the relative magnitudes of the A_i's to be $A_1 = 1$; $A_2 = .933$; $A_3 = .745$; $A_4 = .471$ and therefore the powers must be in the ratio of 1:.870 : .555 : .222.

These results were used in the equation given above for finding the far field pattern. This equation contains the quadratic exponential term to account for the parabolic phase front across a pyramidal horn aperture. The integrals were evaluated on a computer using two values of the parabolic phase - a half radian and a full radian variation. At 35 GHz for the 16λ total aperture, and a seven element array, the half radian case corresponds to a horn whose axial length is 2.174". The full radian variation would correspond to a horn of 1.057" axial length.

The patterns for these two cases were calculated and some are included here. Figure 2 shows the H-plane pattern for the case of a 1/2 radian variation in the phase front, i.e. the longer horn. The region of primary interest is of course from -12^o to $+12^o$



$$F(x) = A + B\cos\frac{\pi x}{L}$$

FIG. 1—EQUAL AREA APPROXIMATION TO RAISED COSINE DISTRIBUTION

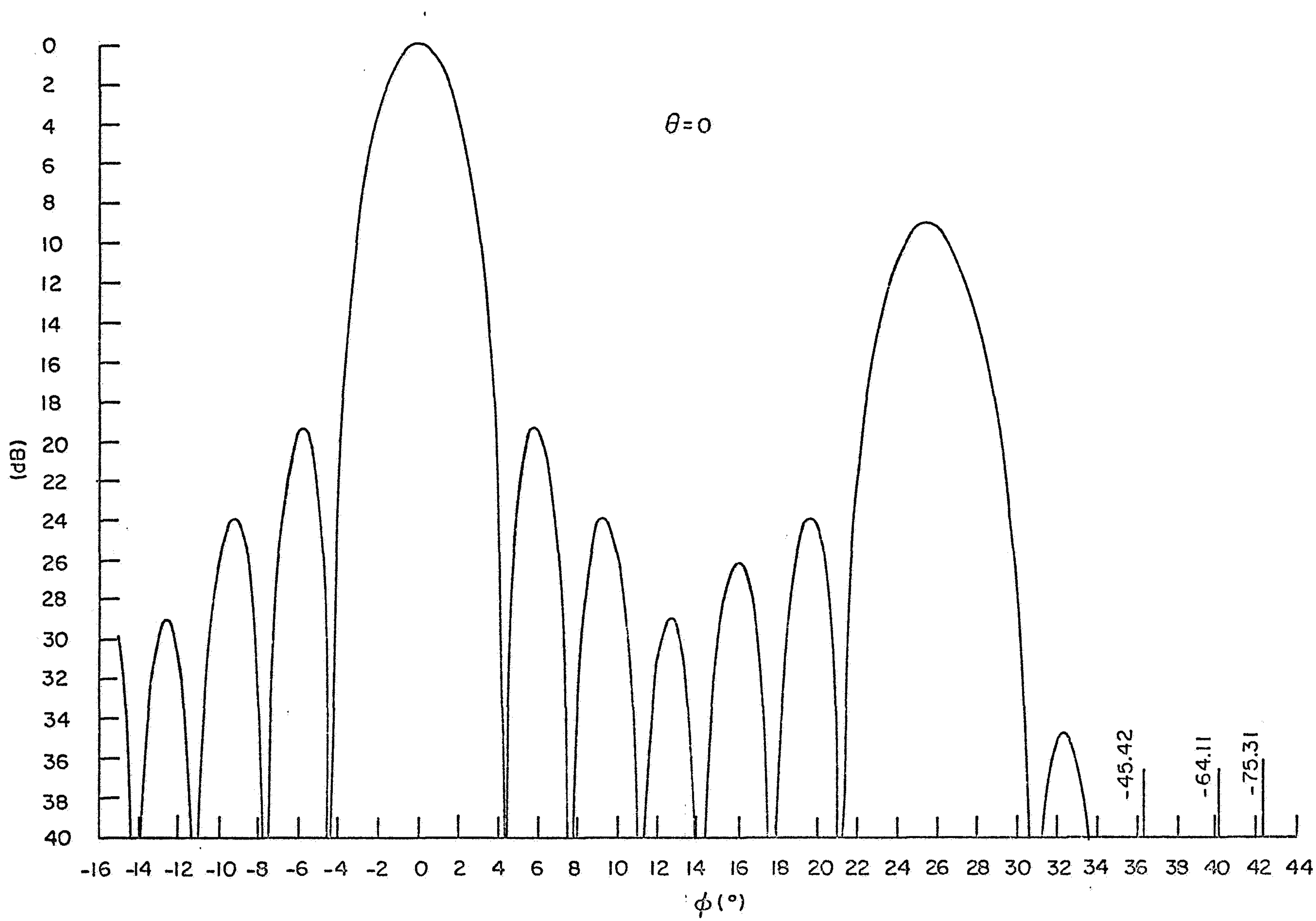


FIG. 2 - ELECTRIC FIELD CALCULATED WITH VARIATION IN AMPLITUDE TERM.

for the synchronous orbit case with allowance for errors of positioning in the satellite. This plot is for the 10 dB illumination taper across the aperture. The grating lobe appears properly at 26° and well removed from the region of interest.

Figure 3 presents the same case steered 10° off boresight. The main beam has deteriorated about 1 1/4 dB. The grating lobe to the opposite side of boresight is now enhanced by the element pattern, but in the angular region of interest is still down some 15 dB.

For the opposite polarity the unsteered beam is, for all practical purposes, the same as Figure 2. In the extreme steering condition (see Figure 4), some obvious differences occur in the far sidelobe pattern. In the region of interest, however, the patterns can be almost exactly overlaid.

The computer results also indicated that the difference between the shorter horns and the longer horns was an increase in loss which was about .7 dB in the worst case. However, such small losses can accumulate and the sum effect could be serious so it was decided to choose the longer horn (1/2 radian phase variation) even though this added an inch to the length of the system. Since it is not possible to have a square pyramidal horn on rectangular guide with equal phase centers in both planes, the horns were designed to have slightly more than the 1/2 radian variation in one plane and slightly less in the other.

In addition it was decided that the performance of the 7 x 7 element array was acceptable and that there was no need to investigate higher orders. The final design of the output array, then, was a 7 x 7 matrix of uniform horns, equally spaced and completely filling a 16λ square aperture. This array is shown in Figure 5.

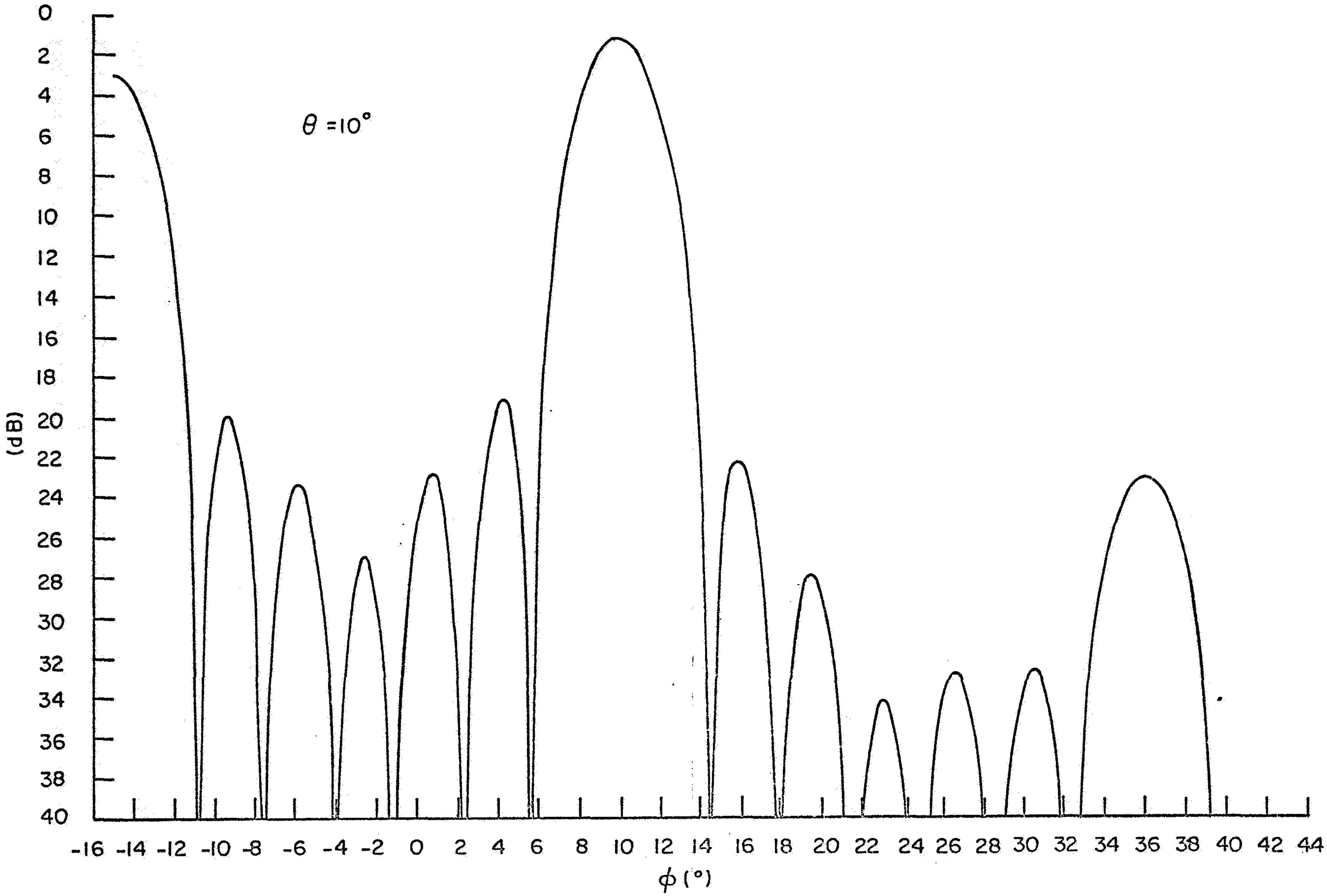


FIG. 3 - ELECTRIC FIELD CALCULATED WITH VARIATION IN AMPLITUDE TERM

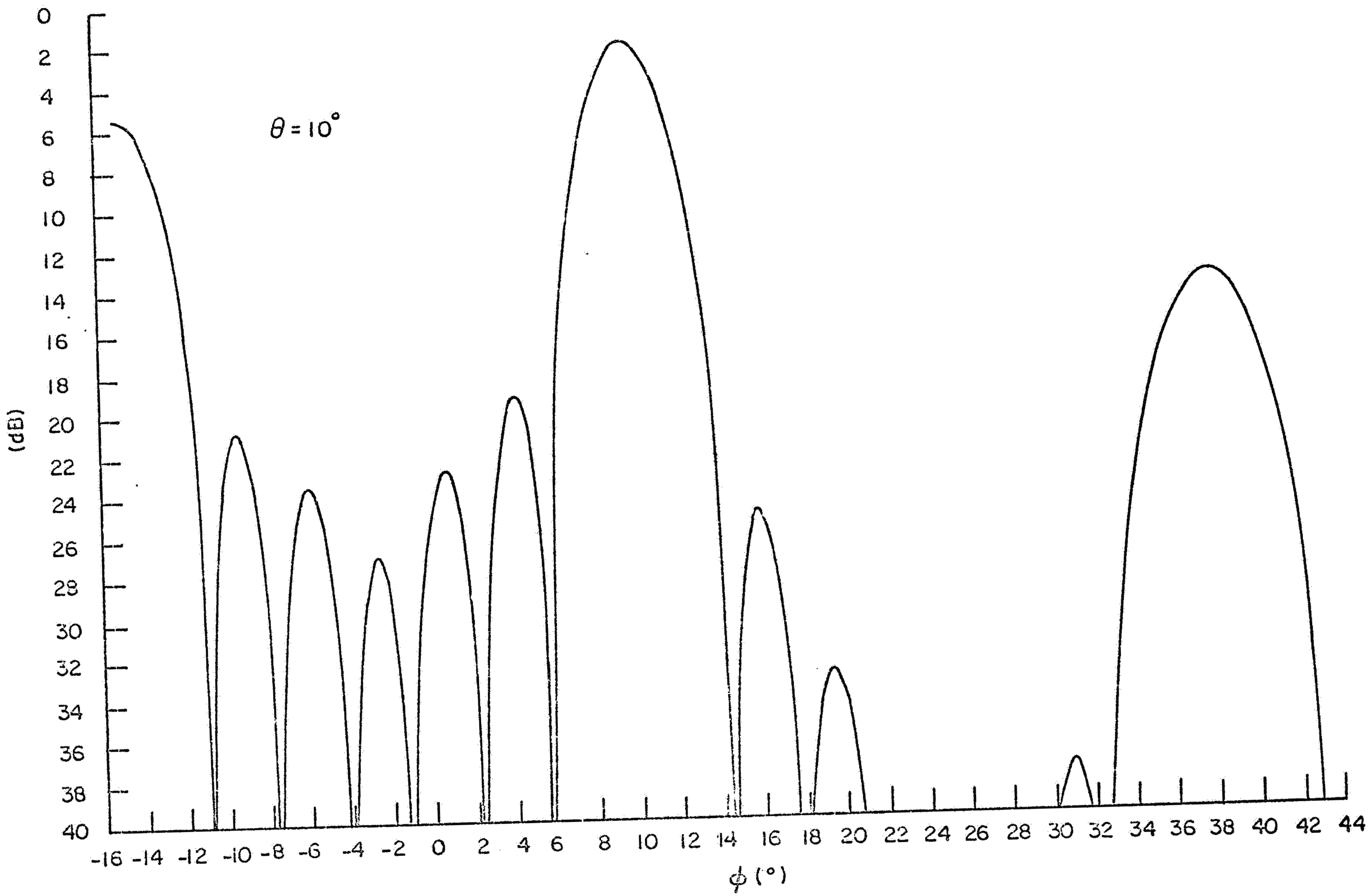


FIG.4-ELECTRIC FIELD CALCULATED WITHOUT VARIATION IN AMPLITUDE TERM

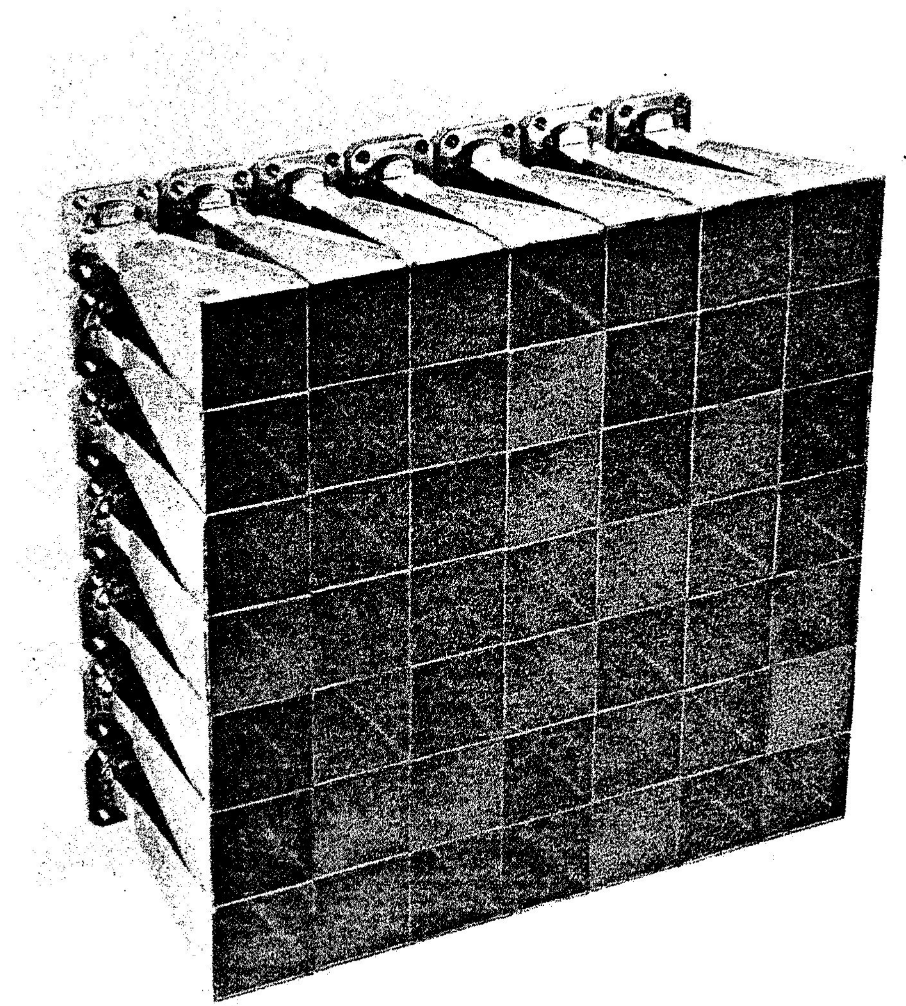


Figure 5 - Feasibility Model of the Uniform Output Array for 35 GHz

6.2 Steering Mechanisms

6.2.1 Up-Conversion Method

In Section 2.1.3 the recommendation was made that the most practical up-conversion technique to pursue was the one which injected the phase control at the fundamental frequency since for this case, there is less of a component restriction. The energy is then multiplied and the harmonic channeled directly to the output array.

For this type of design it becomes essential to know the relationships between the phase of the output signal and the phase and amplitude of the input signal. (The amplitude enters in if identical doublers are to be used and an amplitude taper is desired across the array.)

In order to investigate experimentally these phase transfer characteristics, a varactor doubler from 17.5 to 35 GHz was designed and a pair of these units fabricated. The units are pictured in Figure 6. The input waveguide is RG-91/U and provision is made in this guide for tuning screws. Past the tuning screws is a dumbbell type resonant iris across the guide. This iris is located midway between the rectangular block which accepts the tuning screws and the square block in which the varactor is centered. The fundamental waveguide is terminated in a short circuit at the varactor position, where the harmonic waveguide begins (RG-97/U). A semicircular cutout is made here to permit the varactor package to be located at the plane of the short circuit. The distance from the iris to the shorting plane is slightly more than a half guide wavelength at the fundamental.

The insertion of the varactor itself into the waveguide is adjustable, and there is a tuning screw which is coaxial with the varactor and adjustable from the opposite side. The output is also

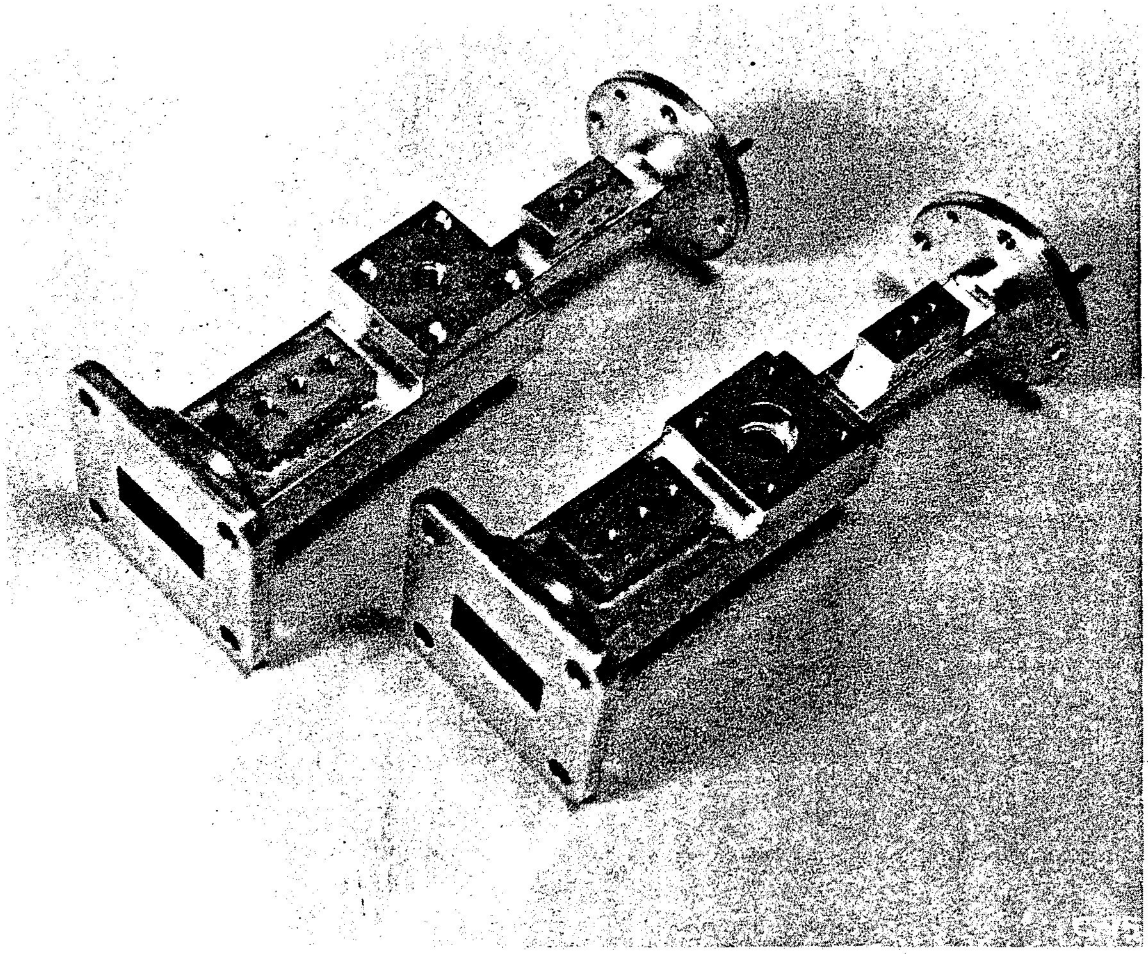


Figure 6 - 17.5 to 35 GHz Varactor Doublers

provided with a three screw tuner for matching.

To test these devices a phase bridge was set up. Each multiplier was then positioned such that the harmonic output fed one leg of the bridge network. The input power to the two was then lowered by use of a single precision attenuator preceeding the fundamental power divider. This was done to simulate the action which might occur in a harmonic transmitting array should the primary source begin to lose power.

Since the response of this type of device is dependent on input power a great deal of effort was extended to producing two units with matching efficiency characteristics over a fairly large input power variation. Success was finally achieved in almost exactly matching two units over the first two dB variation and having no more than 5% difference in efficiency over the next 3 dB. These data are plotted in Figure 7.

Next the characteristics were compared in a phase bridge with the power input being reduced simultaneously to each. The results are plotted in Figure 8 and are quite disappointing. For a 3 dB change in the input there occurred over 100° difference in the output phases. Over the interval where the efficiency characteristics matched well, there is still a steady deviation. Although this effect is largely unexplained it was postulated that an improvement could be made by redesigning the varactor mount to make it a lower Q device. Since the resonance curve would then be flatter, changes in power should effect smaller changes in phase. However, the redesigning would have required a great deal of time and effort and therefore, since this technique can be used only in the transmit mode anyway, it was decided that a different approach would be more advantageous.

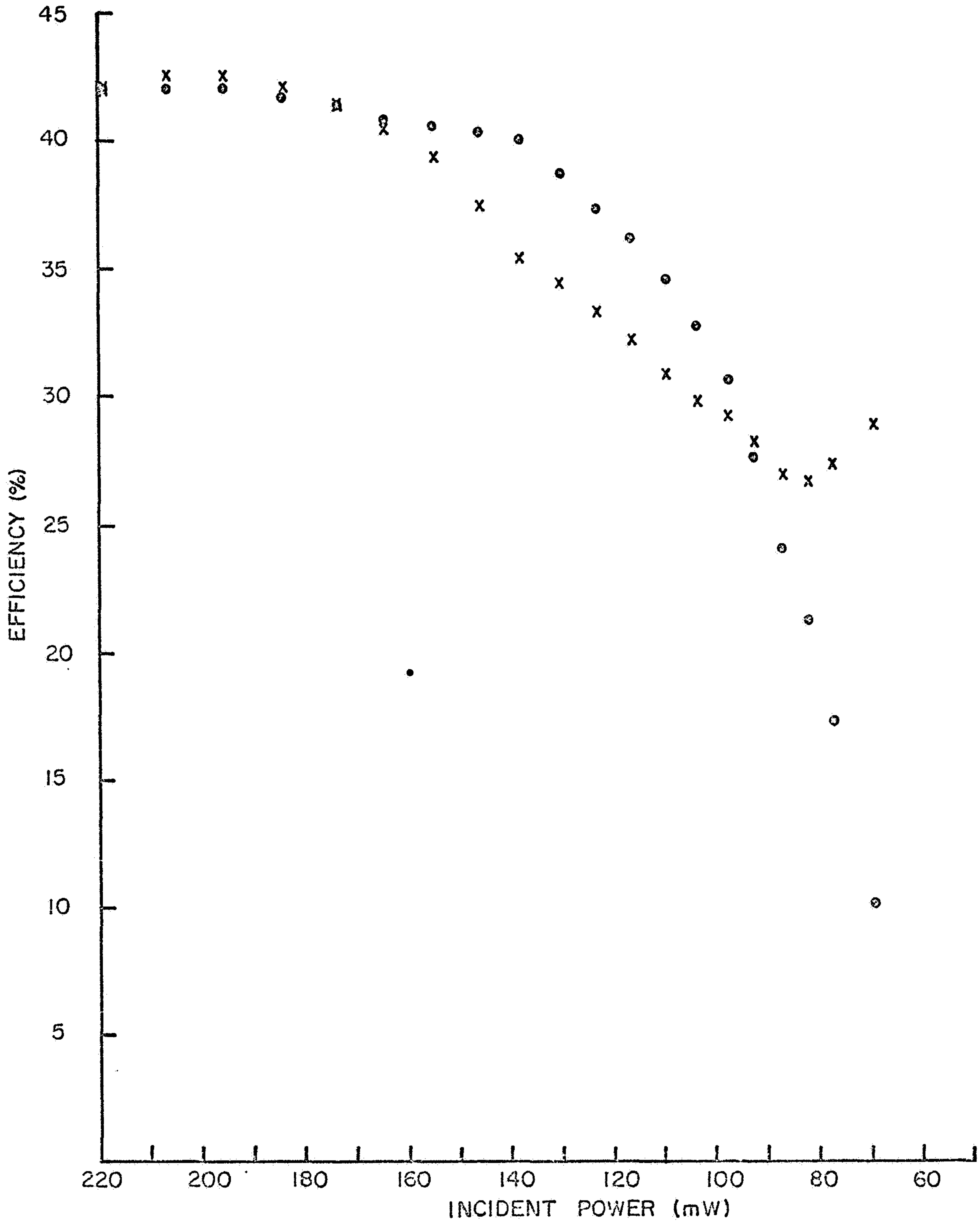


FIG. 7- EFFICIENCY CHARACTERISTICS OF TWO MULTIPLIERS

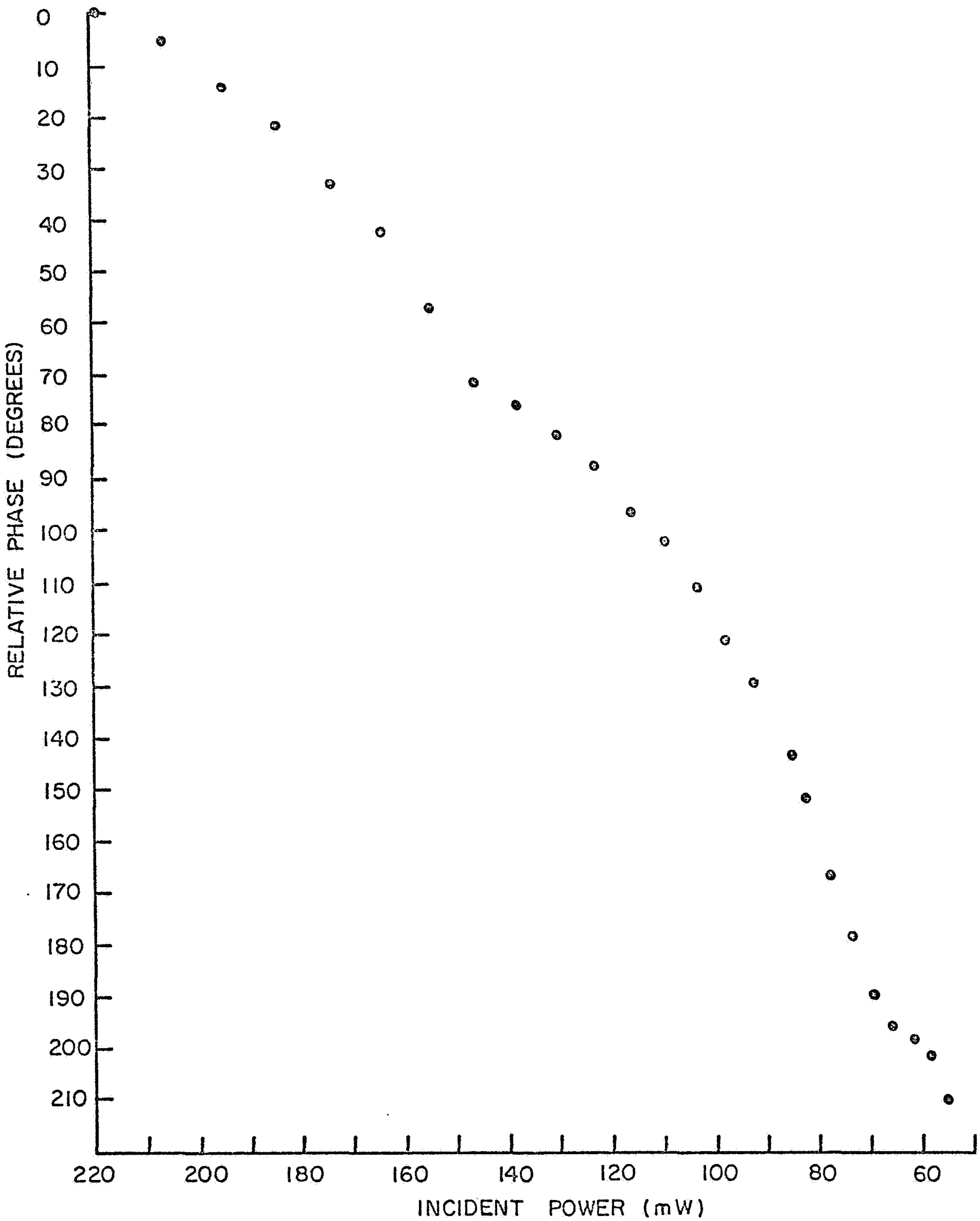


FIG. 8-PHASE DIFFERENCE IN OUTPUT OF TWO MULTIPLIERS

6.2.2 Retrodirective Systems

When one is comtemplating very accurate steering of a large, narrow beamwidth, array antenna, retrodirective techniques have much to offer. Here one would be concerned with the random and unpredictable effects of a turbulent atmosphere which could interfere with pointing accuracy significantly. The technique of phasing the antenna by transmitting the conjugate of a wave front which has traversed this random medium then serves as continuous compensation as long as the round trip time is short compared with the period of the atmospheric changes. For synchronous altitudes, present data would indicate this condition is well satisfied at 35 GHz.

When one is comtemplating a rather small array, however, and a beamwidth of 3 or 4 degrees, the perturbing effect of the atmosphere becomes much less significant, and the additional complications attendant to retrodirective techniques are not well compensated by additional benefits. Consider some of the complications which do occur.

There are many variations possible on the retrodirective approach but there is one problem common to all which is basic and a major detraction from their application here. One of the first operations which must be performed in the retrodirective system is the mixing of an incoming pilot signal with a pump signal at approximately twice the frequency of the pilot. This in turn means that one or the other of these two signals must be at an adequate level to serve as local oscillator for the mixer.

It can be quickly shown that the pilot signal cannot provide this power, hence it must be available on the satellite. Now consider a 49 element array at 94 GHz. One must furnish a signal in the milliwatt level at 186 GHz to each of 49 mixers, and do this precisely in phase. This calls for a better multiplier chain than might be contemplated for the complete transmitter using the techniques described earlier, and can be rejected on this basis. The situation is somewhat relaxed at 35 GHz where the pump power need only be at say 69 GHz, but it is nonetheless a fairly large requirement.

In spite of the generation of the millimeter power for mixing, the actual transmitter power will still be quite low unless amplifiers are incorporated at each element. This statement can be supported by a brief calculation. Being quite liberal with the available resources on the ground, postualting a 70 dB gain antenna with 100 kW power output leads to a power density, at synchronous range, of about 6.2×10^{-9} watts/cm^2 (neglecting atmospheric losses). For the 16λ square aperture then the power intercepted is at most 1.2 μW. Since essentially no gain will be realized in the varactor conversion, this, after the information is added, becomes the transmitter power which is entirely inadequate.

Hence for this small aperture case, the extra weight, power and complexity necessary to implement the system is not rewarded sufficiently by improved performance and the system was not considered further. Variations have been investigated wherein one alters the approach to the point that the pilot signal is used to set up a self phasing array and these then combined with a transmitter lead to equally - or even more- complex arrangements and were similarly discarded.

6.2.3 Phase Shifter Techniques

Since the recommended systems are predicated upon the use of phase shifters in each element, it becomes obligatory to show that such systems are indeed practical at millimeter wavelengths. The phase shifters in general use at microwave frequencies are usually either semiconductor or ferrite types. Within each group, there are sub-classes of analog and digitally controlled types. As of this moment, neither category (semi-conductor or ferrite) could be considered as an off-the-shelf item at millimeter wavelengths, although the ferrite devices are probably closer in this regard. This study then recommended that ferrites be used in the final application, and the reasoning behind this is summarized below.

6.2.3.1 Semiconductor Phase Shifters

One of the strong selling points for varactor phase shifters is their capability for rapid switching. This, however, is not one of the requirements for the present application, and hence some of the limitations of a diode device become more apparent and significant.

Some insight can be obtained as to the working potential of the diode for millimeter wavelength by examining the devices available at microwaves. If we assume that the varactor cut-off frequency would be scaled in proportion with the frequency scaling, then some parameters such as insertion loss would remain approximately constant. Other parameters, however, such as power handling capability will change due to the reduced junction areas. Hence, consider a commercially available phase shifter at 9 GHz which has a maximum phase shift of 180°, average power handling capability of 4 watts and an insertion loss of 1 dB. To scale this in frequency by a factor of 10 while retaining the insertion loss characteristics would significantly reduce the power capability and hence limit its potential as applied to transmitting arrays. Furthermore, this only provides half the phase capability necessary.

For the millimeter arrays, one would not be inclined to use the reflecting type of phase shifter which necessitates the use of an isolator in conjunction with it. The transmission type with the diode in the waveguide has the fundamental theoretical limitation of a 90° phase shift capability per diode, and a practical limitation somewhat less than this. Hence, each output waveguide would require at least four of these diodes. The waveguide circuit for this configuration presents impedance matching problems in addition to the development of the diode itself.

If one specifically needed the fractional micro-second switching capabilities and wanted scanning capabilities at Gigacycle rates, the emphasis here would be the urging of diode development work. Since this development work has not been begun at millimeter wavelengths, and rapid switching is not a requirement, then the case for diode phase shifters is not strong at this time.

6.2.3.2 Ferrite Phase Shifters

Unlike the semiconductor case, development of ferrite phase shifters is currently in progress at both 35 and 94 GHz, and in both analog and digital types. The analog type are the conventional Reggia-Spencer phase shifter and are specified in at least one catalog (TRG, Inc.) to provide a minimum of 360° phase shift with a maximum insertion loss of 1.2 dB at 35 GHz and 1.8 dB at 94 GHz. Bandwidth capabilities on these units are 5% which is entirely adequate for the present application. The size of the current package is somewhat large for convenient use in an array, but this could be reduced. The only real argument against analog phase shifters is the drive power required. If one expects the array application to be one of steering to a given beam position and holding this beam position for long periods, latching ferrite devices are extremely well suited. The phase shifters cited above require 100 ma of current in a $16\,\Omega$ coil for 360° phase shift, i.e. .16 watts. If one were considering a 49 element array wherein the broadside beam required no phase shift and then steered this beam 5° in one plane, one can calculate that with these phase shifters the drive power required for holding would be two and a half watts. If power drains of this nature are not essential, then the satellite application will justify their elimination.

The power drain problem is alleviated by the use of latching ferrite phase shifters. With these in the system the beam remains in position, whether steered or unsteered, with no further power drain. The other projected parameters were sufficiently good to warrant their use in the feasibility model. One parameter, which is more of a goal if not actually realizable, should be noted here. This is the need for constant insertion loss over the range of frequencies of interest. It is easily seen from the above discussion that a change in insertion loss at each element would affect the ratios in the power distribution which in turn would degrade the pattern. However, the state-of-the-art of these components requires that this remain only a goal.

There was one item which remained unspecified at this point and this was the number of bits each phase shifter would have. With a three bit phase shifter the maximum phase error at any element would be 45° with an average error of 22.5°. The four bit model would halve these values and the average phase error would be 11.25°. The choice was to be made by programming an error analysis to determine their effect on the far field patterns but at the time the phase shifters had to be ordered this information was not available. This being the case the 4 bit model was chosen and Westinghouse of Baltimore was contracted by NASA to provide these units.

These digital latching phase shifters

have the added advantage of ease of control. They possess the quality of having a natural interface with the satellite's on board computer. If this is not practical a separate control system can be designed to occupy a minimum area using presently available integrated circuits. A proposal has been submitted to the contracting agency by ADTEC to build such a control package as an extension to this contract.

6.3 Feed Design

The foregoing has established some parameters that the feed system, be it conventional waveguide or of a quasi-optical nature, must meet. These include a power split that approximates a cosine-on-a-pedestal distribution along both planes of the output array; low loss; light weight and an output compatible with the phase shifter array. This section describes the development of the recommended systems to meet these requirements.

6.3.1 Conventional Waveguide Networks

There are two basic types of conventional feed systems which can be considered - parallel and series. The former involves the use of a tree of directional couplers or couplers plus hybrid junctions to effect the tapered power distribution. A 49 element array requires 48 such power dividers which could be 24 couplers (with different coupling coefficients) and 24 tees. However, this complexity leads to bulky structures which would be difficult to fit in the profile of the output array. Another fact which reduces their appeal is that they are difficult to fabricate since there is little duplication.

On the other hand, the series method of power division is appealing because it not only can be designed to fit easily into the necessary area but also lends itself readily to duplication of pieces. These identical units could then be economically electroformed using a single design for the mandrel.

One type of series fed structure is a simple waveguide with slots located in the broad wall. By varying the size and/or location of the slot with respect to the centerline of the waveguide, it is possible to achieve a wide range of coupling values. Branching waveguides could be attached such that a tee would be made but with the slot in the main line controlling the coupling into the branch. Several such branches could be assembled along the main waveguide on the same centers (.771") as the output array, with no external mutual coupling effects. Of course there are internal coupling problems present which must be contended with, but these would have to be eliminated or minimized on an experimental basis.

In the light of the above remarks it was decided to begin the feed design using the series coupler method. As noted in a preceding section, the power division must be achieved in predetermined ratios which for a seven port coupler would be .222 : .555 : .870 : 1.0 : .870 : .555 : .222. In order to achieve such a distribution the losses incurred at each port must be considered in order to calculate the amount of power remaining in the guide and successive coupling coefficients. Since this information was not available a test piece was fabricated and the losses due to the junction measured. To calculate the coupling coefficients of the seven ports it was convenient to define the losses incurred over the section from the center line of one branch port to the centerline of the next branch port. Thus for a 7 port coupler there are 6 full sections and 2 half sections - one on either end. For the test pieces the losses were found to be about 4% in traversing a full section and 2% for a half section. Using these values the coupling coefficients were calculated in the following manner.

Let α_i represent the fraction of the power in the main arm which is the coupler at the ith port, and A_i be the output power at that port. Let t_1 be the transmission coefficient through half of a full section, and t_2 the same for the full section. (These will be taken as .98 and .96 respectively on the basis of measurements made.) Then it follows for unit power input that

$$A_1 = \alpha_1 t_1$$

$$A_2 = (t_1 - \alpha_1 t_1)\, \alpha_2 t_2 = \alpha_2 t_1 t_2 (1 - \alpha_1)$$

$$A_3 = (t_1 - \alpha t_1)\, t_2^{\,2} (1 - \alpha_2)\, \alpha_3 = \alpha_3 t_1 t_2^{\,2} (1 - \alpha_1)(1 - \alpha_2)$$

$$\vdots$$

$$A_7 = t_1^{\,2} t_2^{\,5} (1 - \alpha_1) \ldots (1 - \alpha_6)$$

(for the seventh case we are assuming all the power remaining is delivered to the seventh port and hence $\alpha_7 = 1$.)

Now note that these may be written as ratios

$$\frac{A_i}{A_{i+1}} = \frac{\alpha_i}{t(1 - \alpha_i)\, \alpha_{i+1}}$$

where t is taken to be t_1 for i = 6 and t_2 for i = 1, 2, 3, 4, and 5.

Using the previously stated ratios for the step approximation to a 10 dB amplitude taper, one can start from the final coupler and work backwards in the following fashion

$$\frac{A_6}{A_7} = \frac{.555}{.222} = \frac{\alpha_6}{.98(1 - \alpha_6)} \quad \text{or} \quad \alpha_6 = \frac{2.45}{3.45} = .7101$$

The next equation would become

$$\frac{A_5}{A_6} = \frac{.870}{.555} = \frac{\alpha_5}{.96(1 - \alpha_5)\ .7101} \quad \text{or} \quad \alpha_5 = .5165$$

The results of the remaining calculations are summarized in

Figure 9. The loss shown in the schematic is for one branch of the feed system. Since each of these will be fed from a similar power divider to obtain the amplitude taper in the other plane, the total loss for the feed system up to the phase shifters will be about one and a quarter dB assuming no additional losses from some unexpected source.

It was also shown, by calculations too lengthy to include here, that it is theoretically possible to fabricate a three port coupler in the form of a tee with the input matched when the other ports are terminated in matched loads. Hence it is also possible to cascade a series of these as long as the branch arms see a well matched phase shifter, since the input of each junction will serve as the matched termination of the previous junction.

By using the test fixture mentioned above with various sized broad wall slots it was determined that all of the necessary coupling coefficients except one could be achieved. This one, at the sixth port, required that more than half the power remaining in the main line must be coupled out. Clearly, something other than a slot coupler was needed to achieve the coupling with the .771" space requirement. To meet this need a device was designed which might be called a "leaky mitered bend". This consists of a mitered bend with an iris in the mitered wall to allow some power to "leak" through. An evaluation piece was fabricated to examine its practicality and is pictured in Figure 10. The results indicated that the necessary coupling coefficient could be achieved by this method in the space limitation. However, due to the unknown nature of the mutual effects the size of the slots and iris could only be determined experimentally.

This task was begun by designing and fabricating

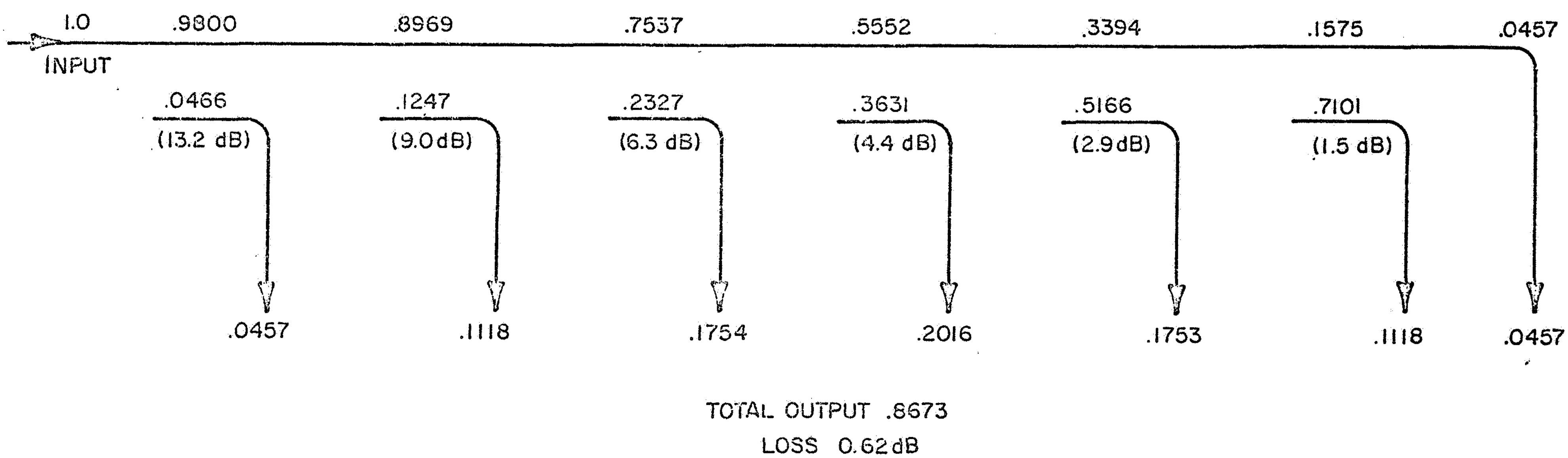


FIG.9-COUPLING COEFFICIENTS FOR SERIES FEED WITH LOSS

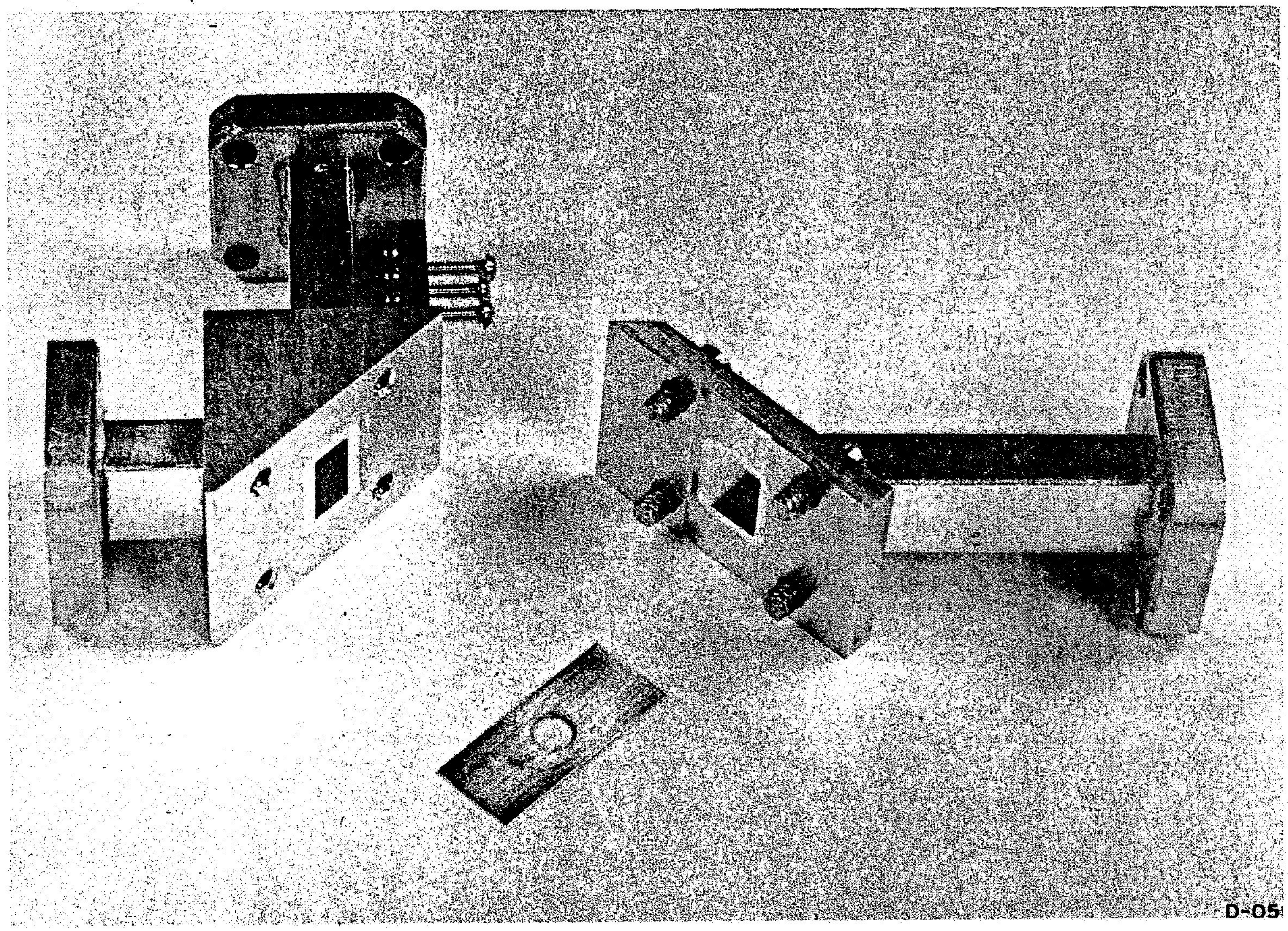


Figure 10 - Test Fixture for Leaky Mitered Bend

a feasibility model for the seven port coupler employing the leaky mitered bend at the sixth port. This device is illustrated in an exploded view in Figure 11. Tuning screws located at each of the first five ports (numbered from the input) enabled each port to be matched out and as it turned out also allowed a limited amount of trimming of the coupling coefficient.

Since the coupling values at ports 5, 6, and 7 were the most difficult to achieve the development of slot size was begun at that end. Once the 7th and 6th ports were determined (by the size of the iris), the 5th port slot was attempted, keeping ports 1 through 4 closed, using various sized slots. Once the proper slot was decided upon and the structure matched to the input, the process was repeated for the 4th slot and so on to the first slot. Although this is simplified from the involved tuning techniques due to mutual effects it is sufficient. The necessary values were achieved within .2 dB at each port. However, losses were higher than the .62 dB predicted for the feed arm and were approximately .8 dB. This implied an overall feed system loss of about 1.6 dB which is still acceptable. The completed feed arm is pictured in Figure 12 and its implementation in the system is illustrated in Figure 13. In the latter figure the units behind the output array of horns represent the phase shifters.

6.3.2 Quasi-Optical Feed System

A parallel development to the conventional waveguide network was a quasi-optical power distribution system. This involved the use of a large illuminating aperture and a 49 element collecting aperture. This section describes the work and analysis done on this approach.

6.3.2.1 Concept and Design

The original concept of the optical

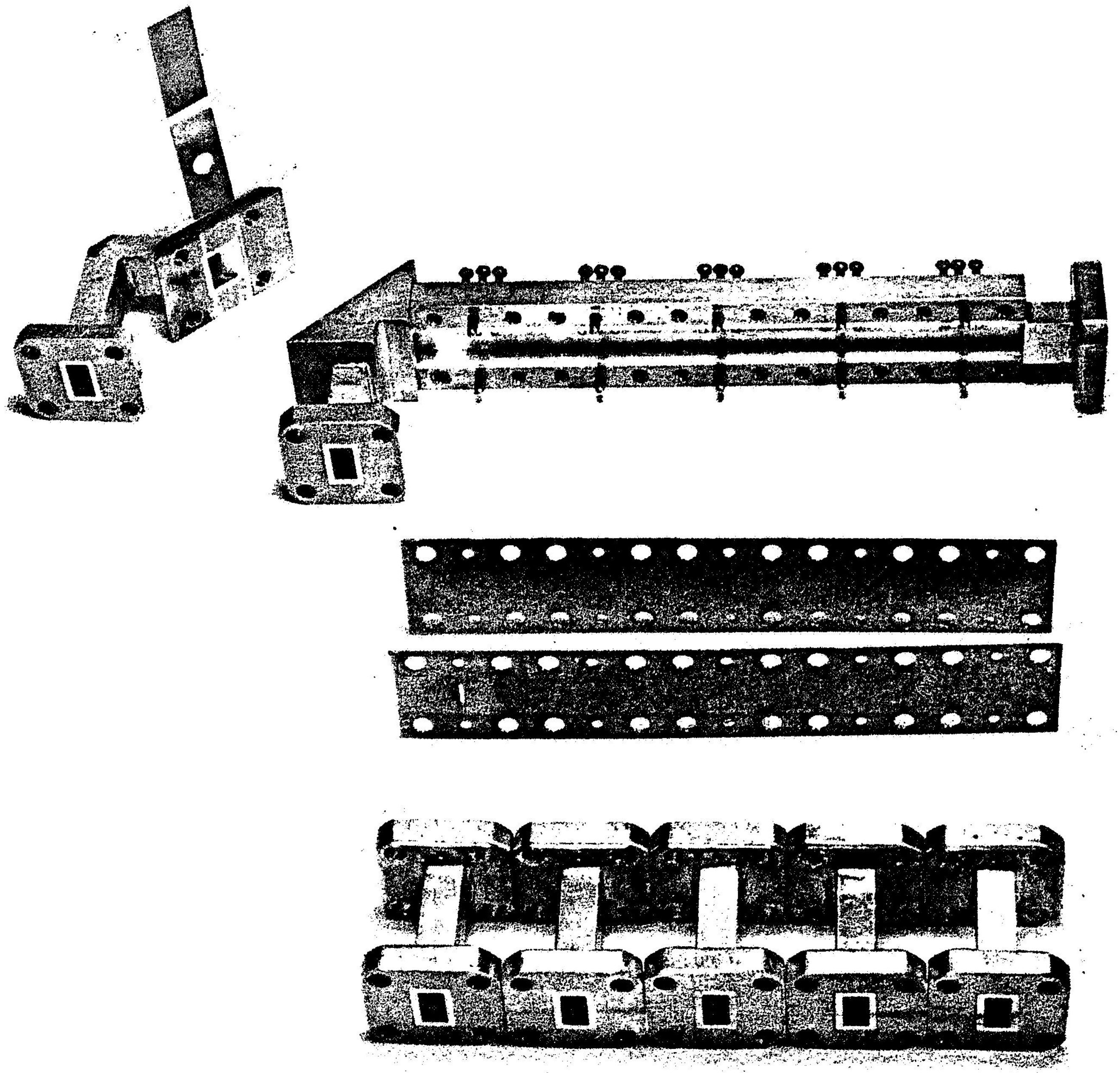


Figure 11 - Exploded View of the Seven Port Coupler Feed

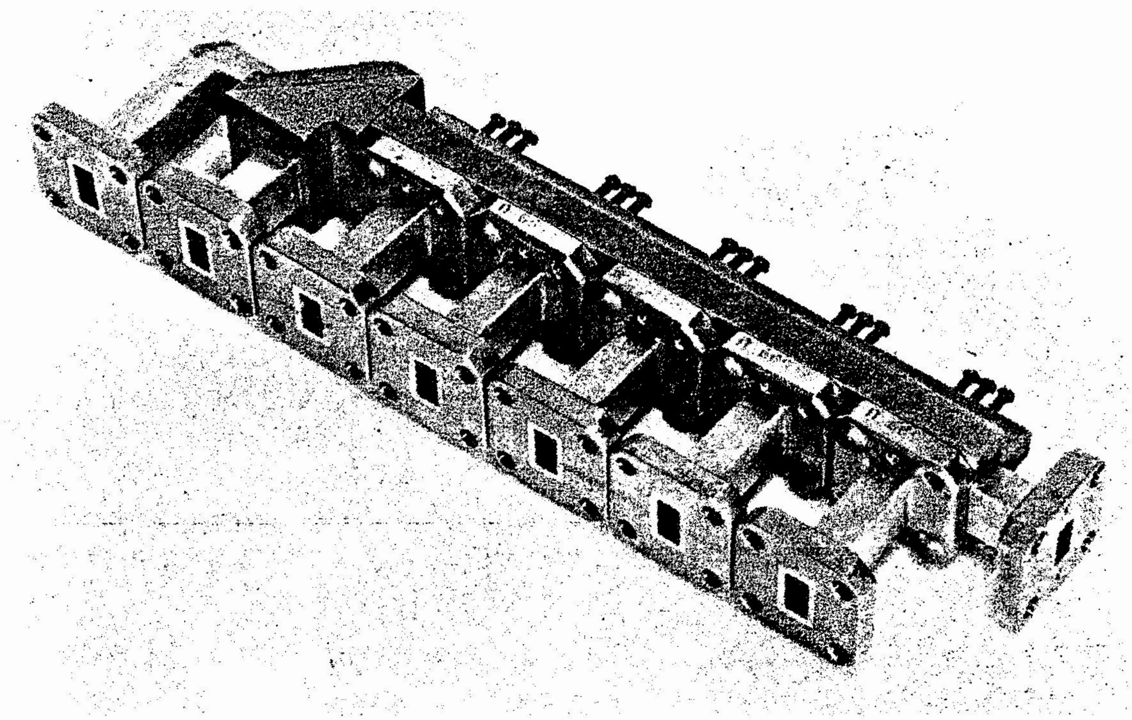


Figure 12 - Completed Seven Port Coupler Feed

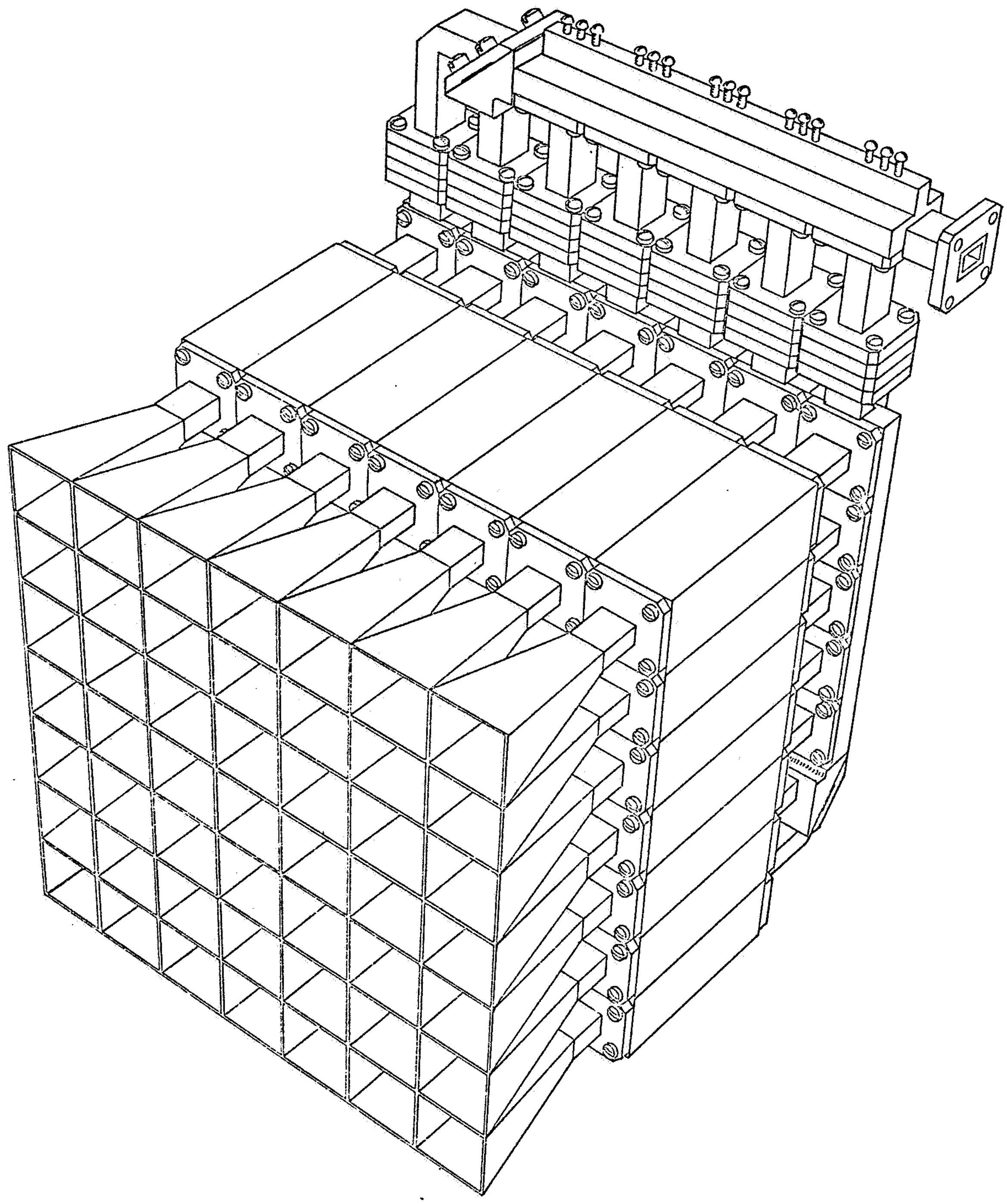


Figure 13 - Feasibility Model of Coupler Fed Antenna System

feed system is illustrated by the drawing of Figure 14. The power which would be radiated by the large illuminating horn is immediately intercepted by the array of smaller horns and brought back into standard waveguide for the phase shifting operation.

It is then radiated from the 7 x 7 output array of pyramidal horns (not shown in the figure). This figure is misleading to the extent that the array of collecting horns is shown to be a uniform array, i.e. all horns identical. However, to effect the proper cosine-on-a-pedestal power distribution these horns would have to vary in size to intercept the correct power. Since the natural power distribution from the illuminating horn follows a cosine squared curve in one plane and is uniform in the other plane, the collecting apertures of the small horns would look something like Figure 15. These aperture sizes were calculated to be exactly those necessary to effect the correct power distribution. But in addition to the amplitude distribution of the illuminating horn, the phase variation must also be considered.

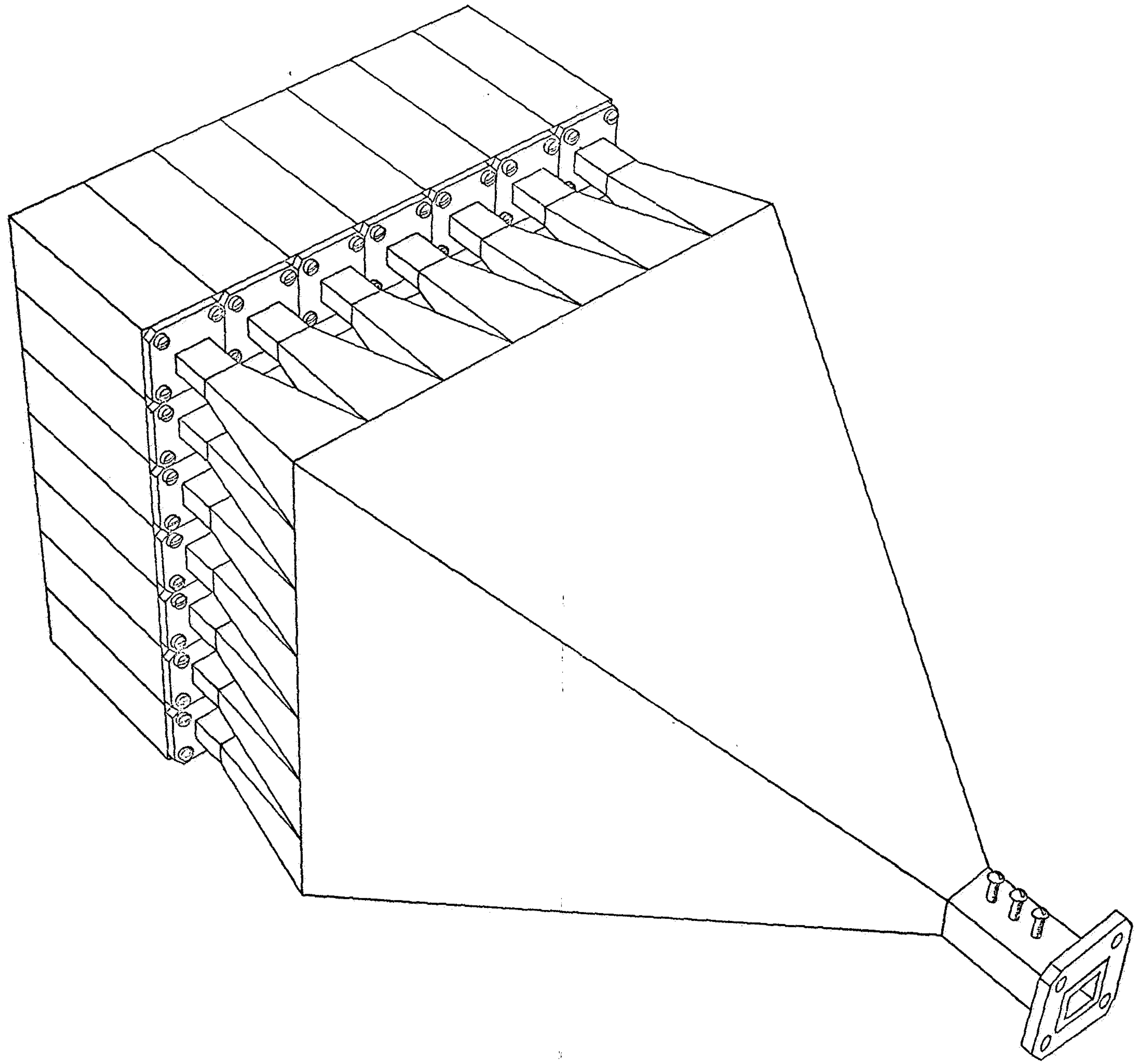
In an unperturbed condition, a standard pyramidal horn of the type depicted in Figure 14 will have a maximum phase variation, T, across the aperture (where T is measured in wavelengths) of

$$T = \frac{L - \left[L^2 - \left(\frac{a}{2}\right)^2\right]^{1/2}}{\lambda}$$

where L = the slant length of the horn to its phase center

a = the width of the aperture

(Note: this is a spherical approximation to the actual parabolic phase variation, but is valid for flare angles of less than 60°.)



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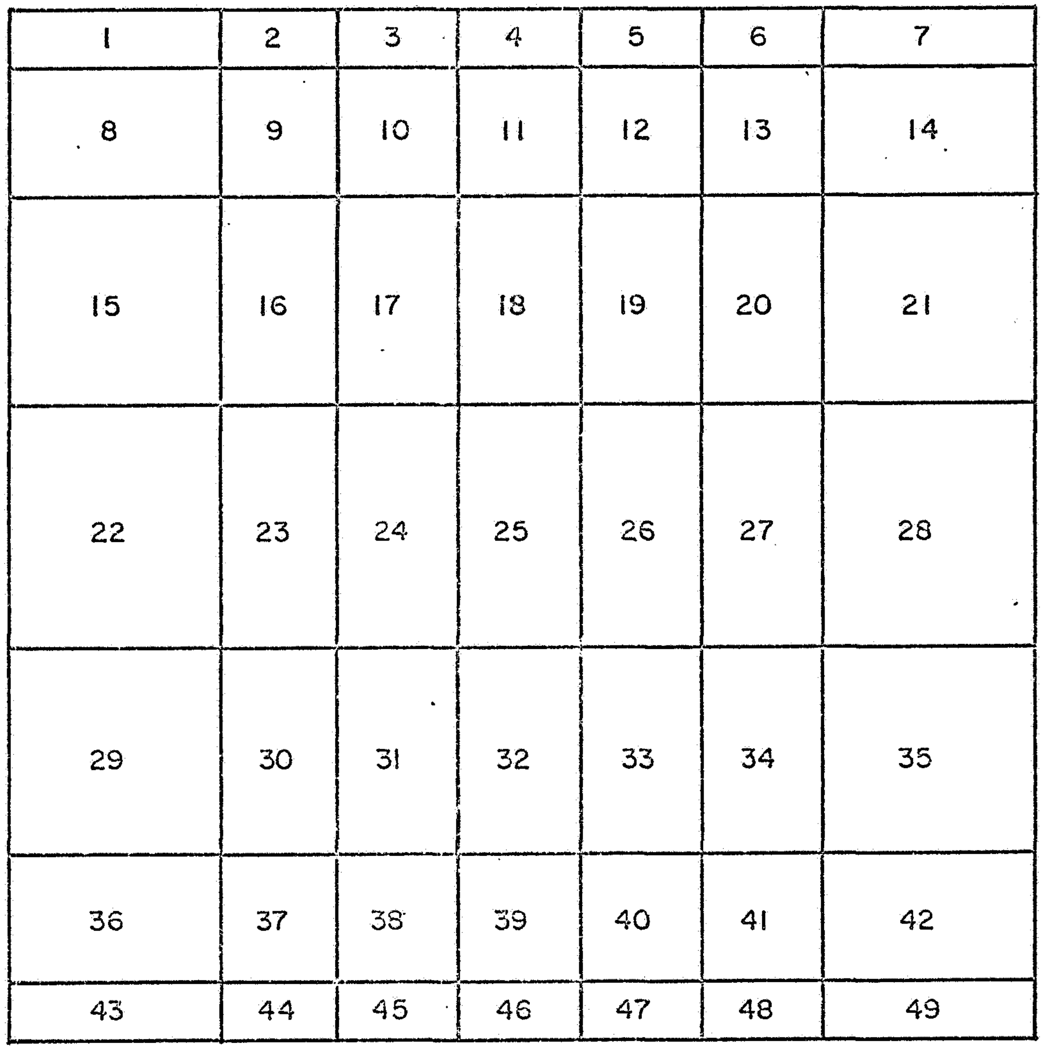

FIG. 15 - CALCULATED APERTURE SIZES FOR A COMPENSATED COLLECTING ARRAY TO EFFECT A COSINE ON A PEDESTAL DISTRIBUTION IN BOTH PLANES.

One must compromise then between the degree of flatness of the phase front desired and the axial length of the horn which can be tolerated. For this development, an axial length of 8" was chosen which results in an uncompensated phase variation of 1.314λ. (We were working under the constraint that the profile of the feed should not exceed that of the 16λ square array. In this case, for convenience in testing they were made precisely equal.)

In order for the collecting horns to be most efficient the phase contour of the illuminating radiation should be the conjugate of that of the collecting apertures. This is virtually impossible. The next best thing would seem to be that of having a uniform mismatch at every horn i.e. a planar phase front. To this end, a phase compensating plano-hyperbolic lens was designed for use with the horn. The surface contour of the lens is determined by the equation

$$\left(n^2 - 1\right) x^2 + 2 f x \left(n - 1\right) y^2 = 0$$

where

f = focal length

n = index of refraction of the lens material

x = dimension parallel to horn axis

y = dimension normal to the horn axis

For this program, the lens was fabricated from Rexolite. The horn-lens combination is illustrated in a exploded view in Figure 16.

Experimental results, Figure 17 to 20 indicated that the lens was perturbing the amplitude pattern somewhat but the phase measurements in Figure 21 show that it was in fact transforming the spherical phase front into a plane one. A VSWR measurement at the input to the optical feed horn showed a fairly large standing wave due to, at least in part, the mismatch at the air-dielectric interface.

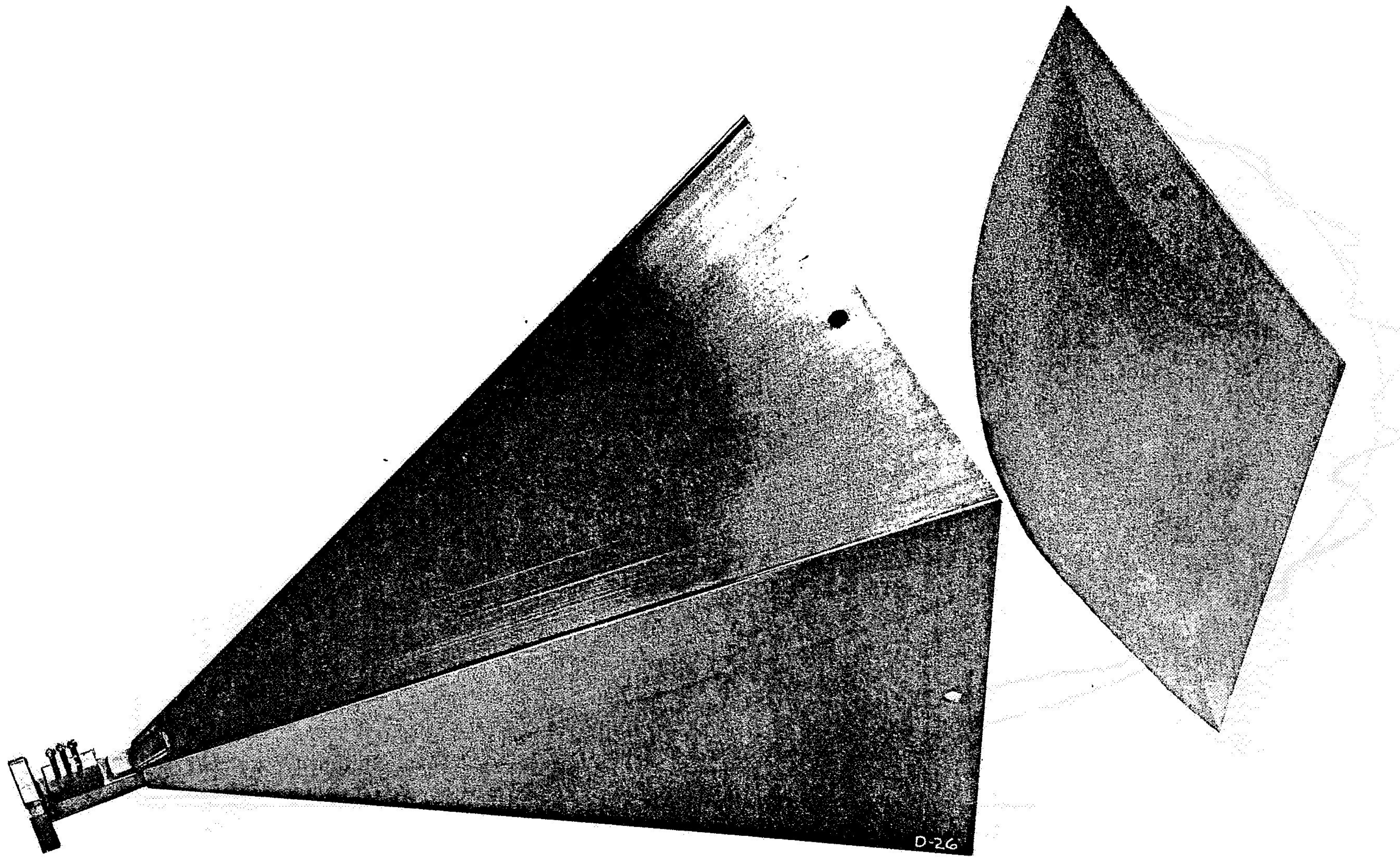

Figure 16 - Exploded View of Optical Feed Horn and Phase Correcting Lens

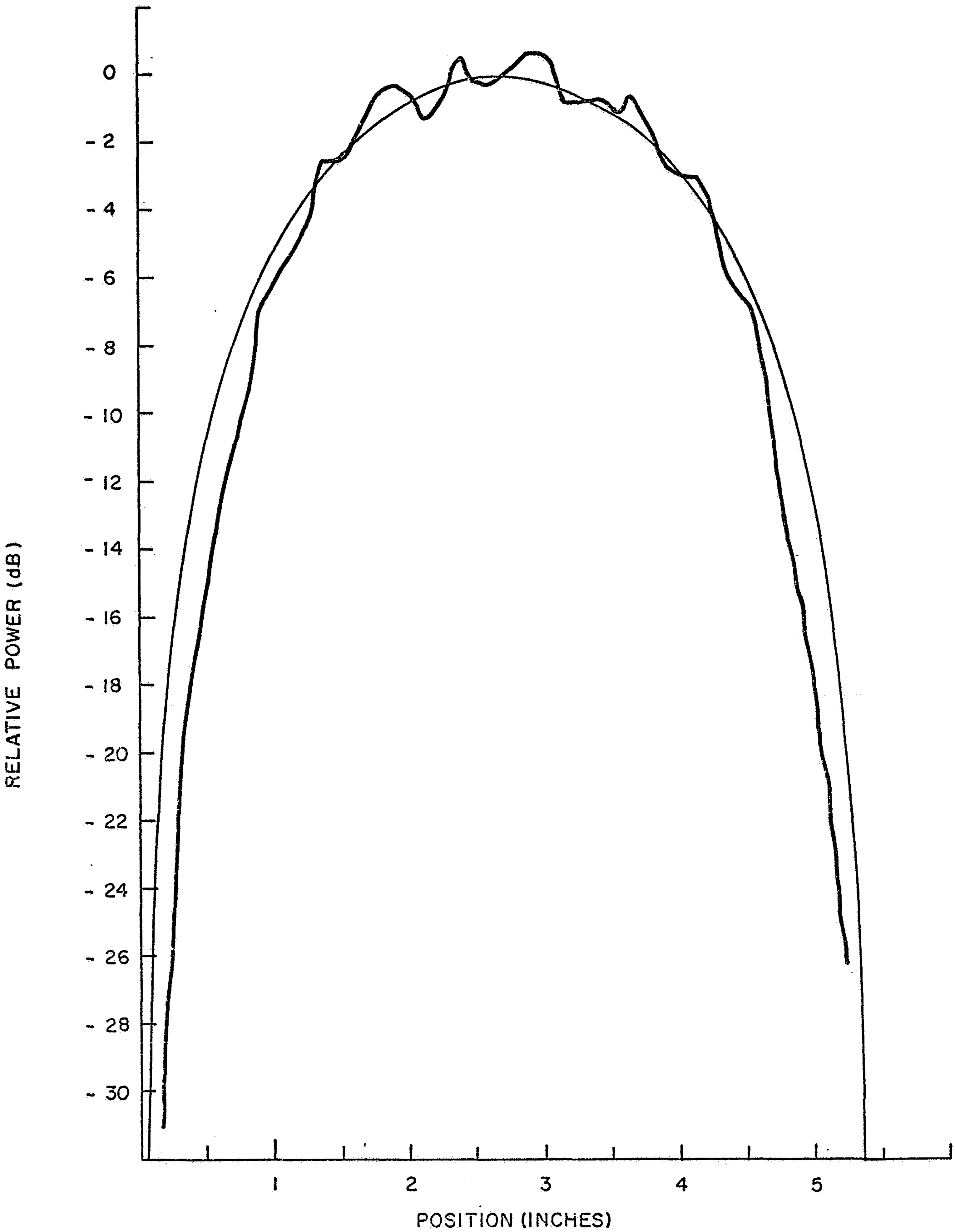

FIG.17- H-PLANE AMPLITUDE (THEORETICAL & MEASURED) WITHOUT LENS IN APERTURE

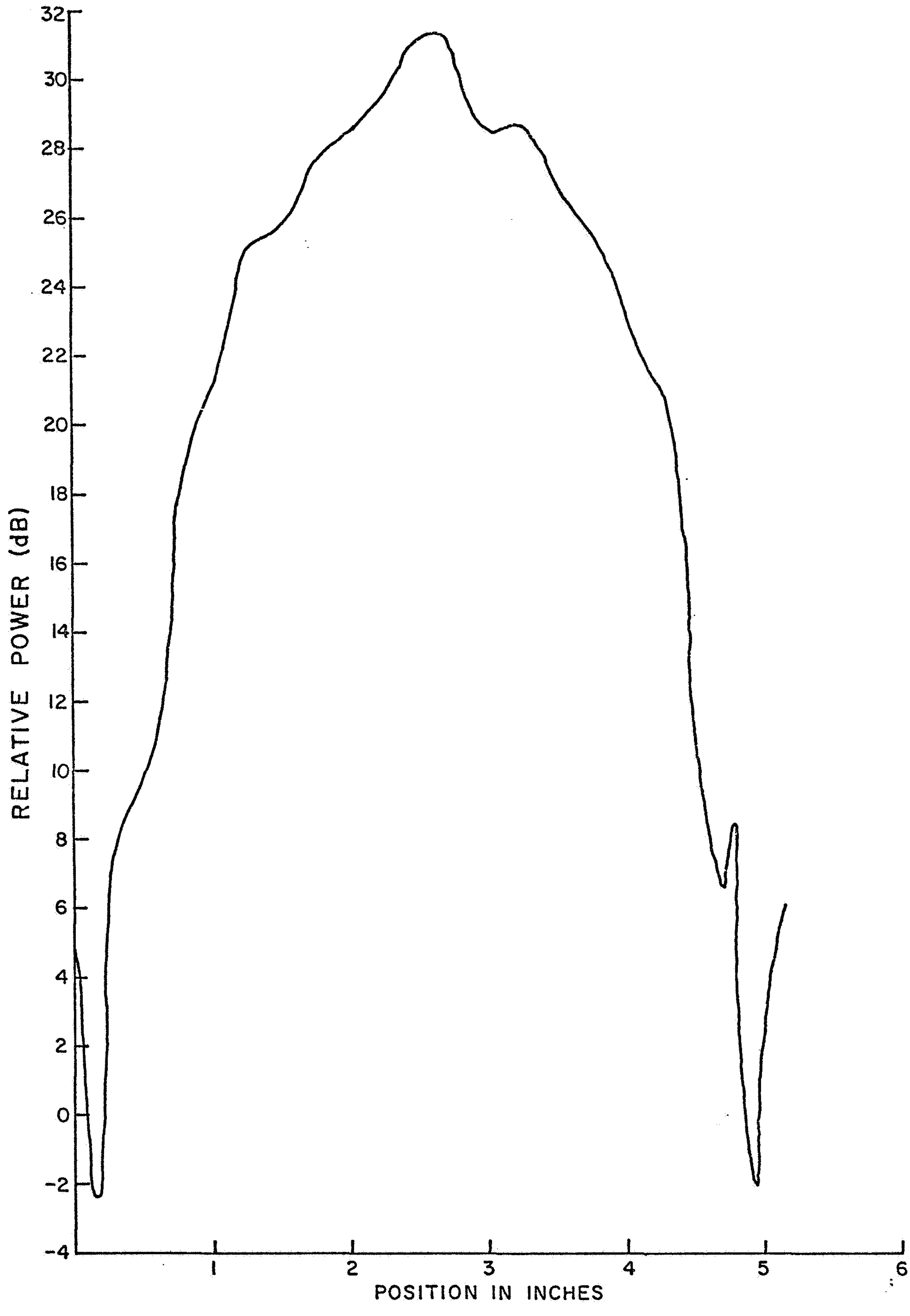

FIG. 18 -RELATIVE POWER vs POSITION (H-PLANE) WITH LENS IN APERTURE

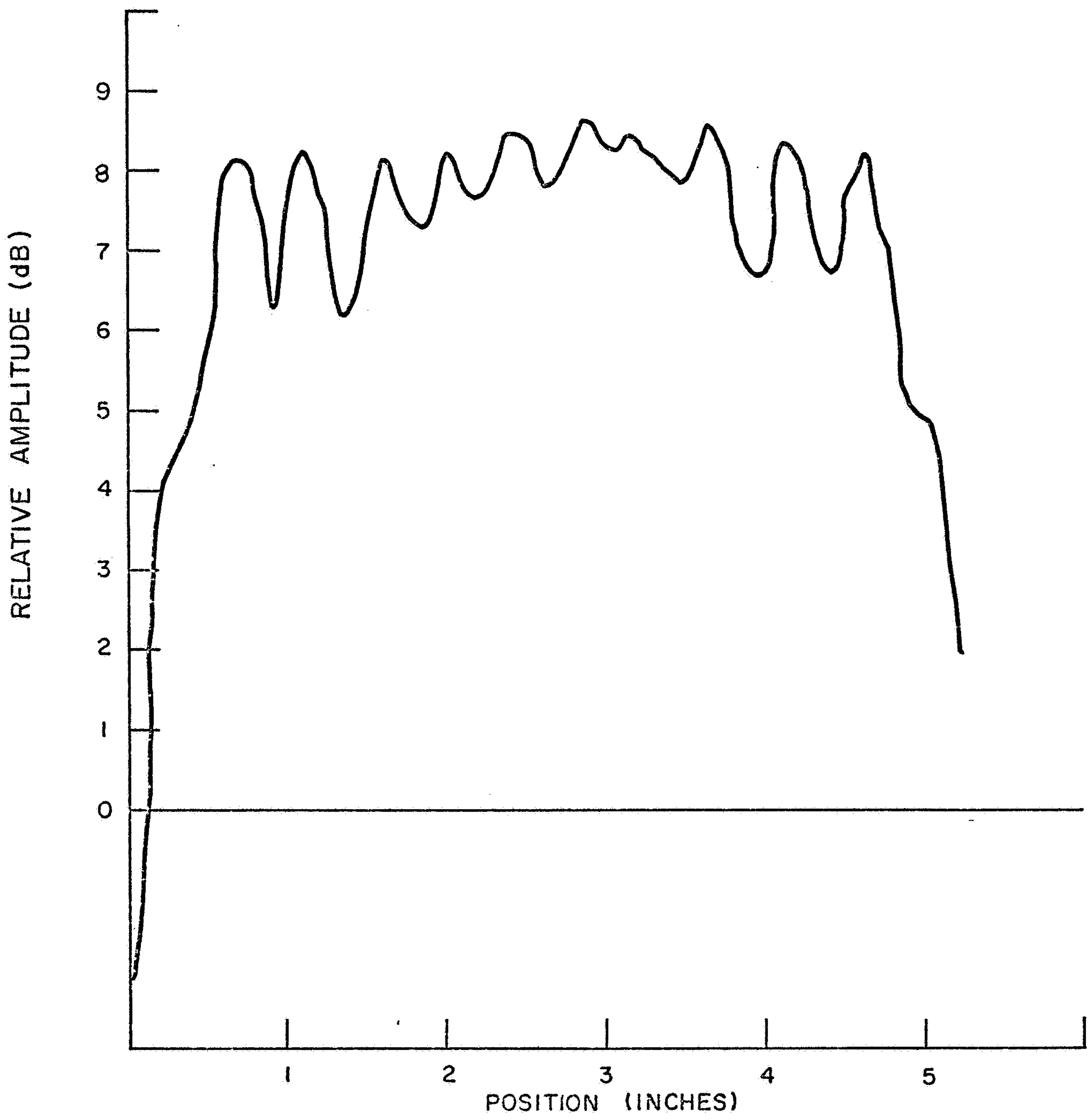

FIG.19 – RELATIVE AMPLITUDE VS POSITION E-PLANE WITH LENS IN APERTURE

FIG.20 -RELATIVE POWER vs POSITION (E-PLANE) WITHOUT LENS IN APERTURE

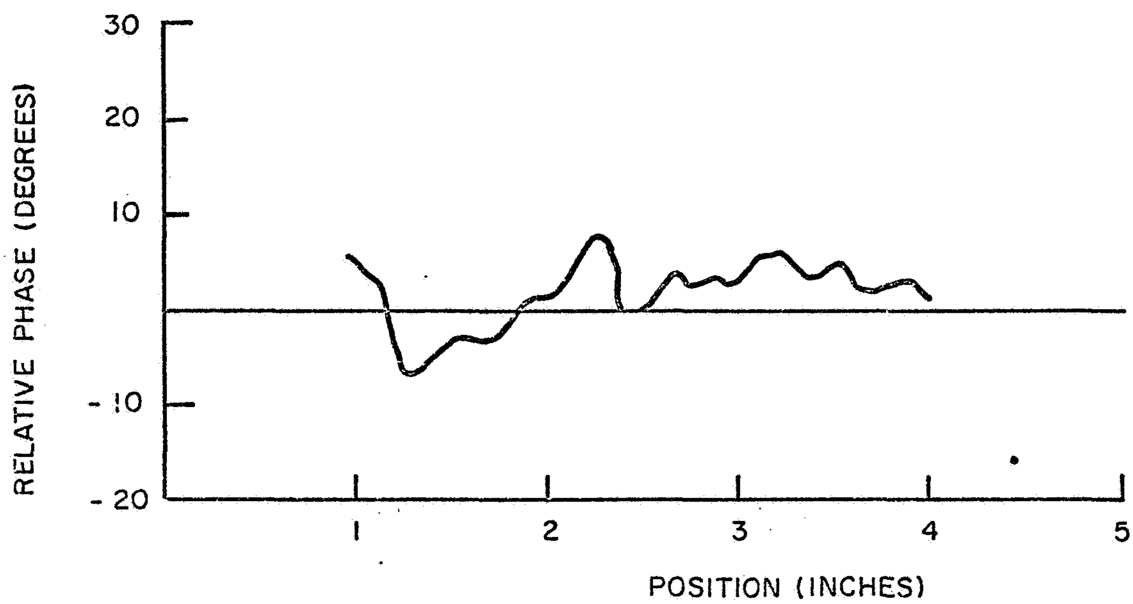

FIG.21 - RELATIVE PHASE VS POSITION H-PLANE

To remedy this situation an effort was made to improve the match at this boundary. There are several methods available but the most practical at this frequency seemed to be the use of a matching layer whose dielectric constant and thickness vary with the angle of incidence to effect a quarter wave layer. Although a continuously varying dielectric constant is not feasible, it has been shown[13] that a layer with a dielectric constant ϵ_m,

$$\epsilon_m = \sqrt{\epsilon_1}\ ,\ \epsilon_1 = \text{dielectric constant of the lens}$$

which is the value for normal incidence, is a good approximation - at least for angles of incidence up to 50°. Then the only varying parameter is the thickness, t, given by

$$t = \frac{\lambda_o}{4\,(\cos\theta)^{1/2}\,(\epsilon_1 - \sin^2\theta)^{1/4}}$$

where θ = angle of incidence, i.e. angle between impinging ray and surface normal.

λ_o = free space wavelength.

The solution of this equation at 35 GHz meant a matching layer .068" thick at the center to .098" thick at the corner of the lens. Since it was required to fit exactly on the hyperbolic contour of the lens, this was a difficult machining job. The VSWR at the input changed from 1.7 : 1 without the coating to 1.16 : 1 with the matching layer. This considerable improvement was not in evidence in other data. Erratic variations from the theoretical powers occured at random in the uniform

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13 E. M. T. Jones and S. B. Cohn, "Surface Matching of Dielectric Lenses", Journal of Applied Physics, 1955, Vol. 26, No. 4, page 452.

collecting array. However, an efficiency measurement indicated that there was only 1.6 dB loss through the system.

Since the correct power distribution was not expected from the uniform horns, an examination of the improvement obtained using amplitude compensated horns was in order. To this end, one row of the above mentioned compensated array (Figure 22) was fabricated and tested. (Note: In the figure the bends shown in Figure 23 on the horns were necessary to preserve the original .771" flange spacing.) It was found that, although there was a considerable improvement in the power distribution, the erratic variations were still present.

Extensive effort was spent trying to isolate the causes of these discrepancies. A test of the system without the lens was tried and this led to an improvement in the regularity of the variations and also reduced the magnitude of them. However, an efficiency measurement without the lens showed losses had increased to 3.7 dB. Another test showed that separating the collecting array from the illuminating horn redistributed the discrepancies and also indicated that a major portion of the problem was due to the reflection of energy from the grid of collecting horns. The test fixture and system are shown in Figure 24.

In an attempt to examine the problem on a smaller scale, a series of sectoral illuminating horns were designed and fabricated. These were not strictly sectoral since they flared in one plane to the size of the collecting horns and in the other plane to the 16 λ dimension of the array. The axial length was kept the same as the original illuminating horn, i.e. 8 inches.

When these structures were tested it became evident that the reflected energy was being trapped in the

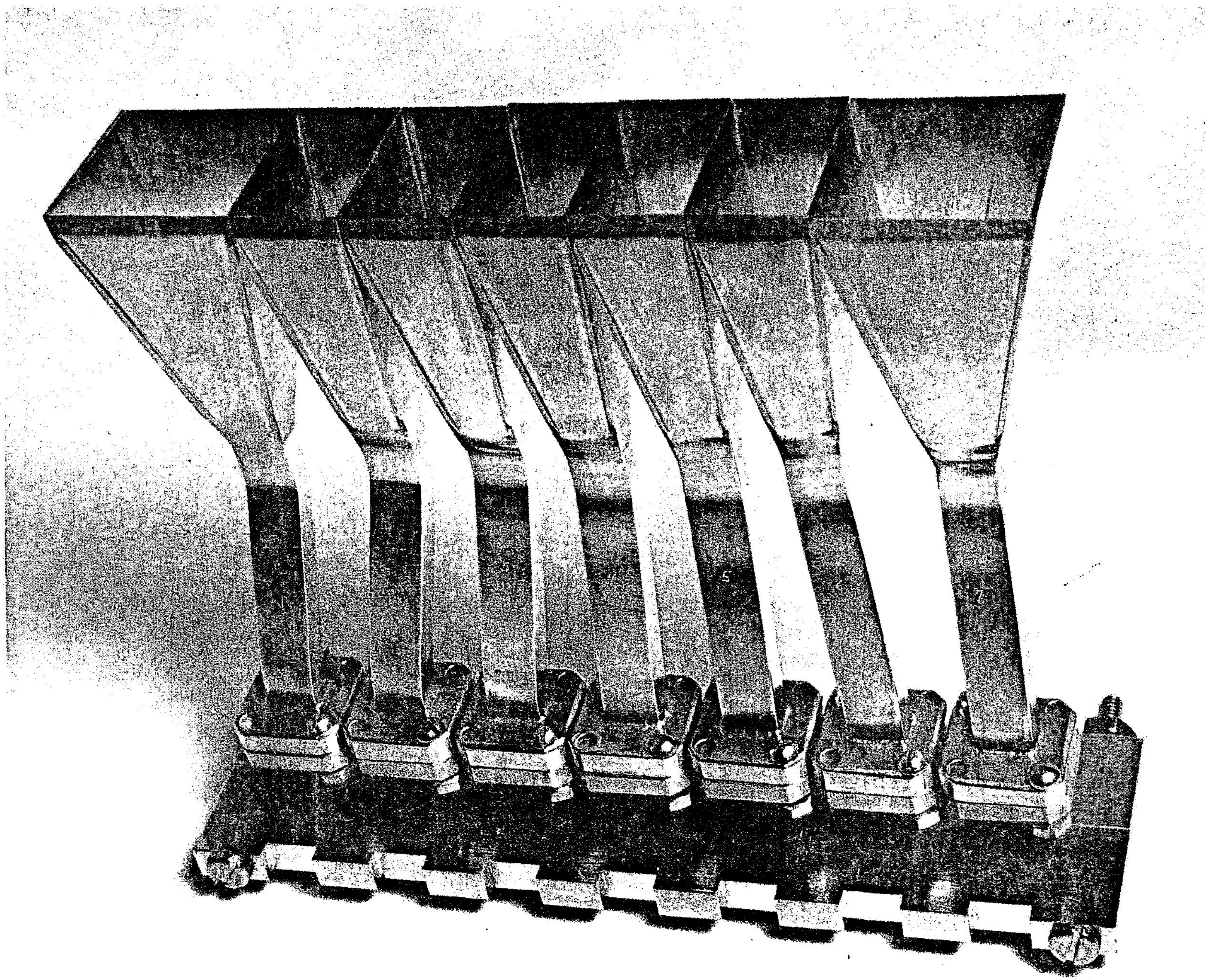


Figure 22 - One Row of Compensated Collecting Horns

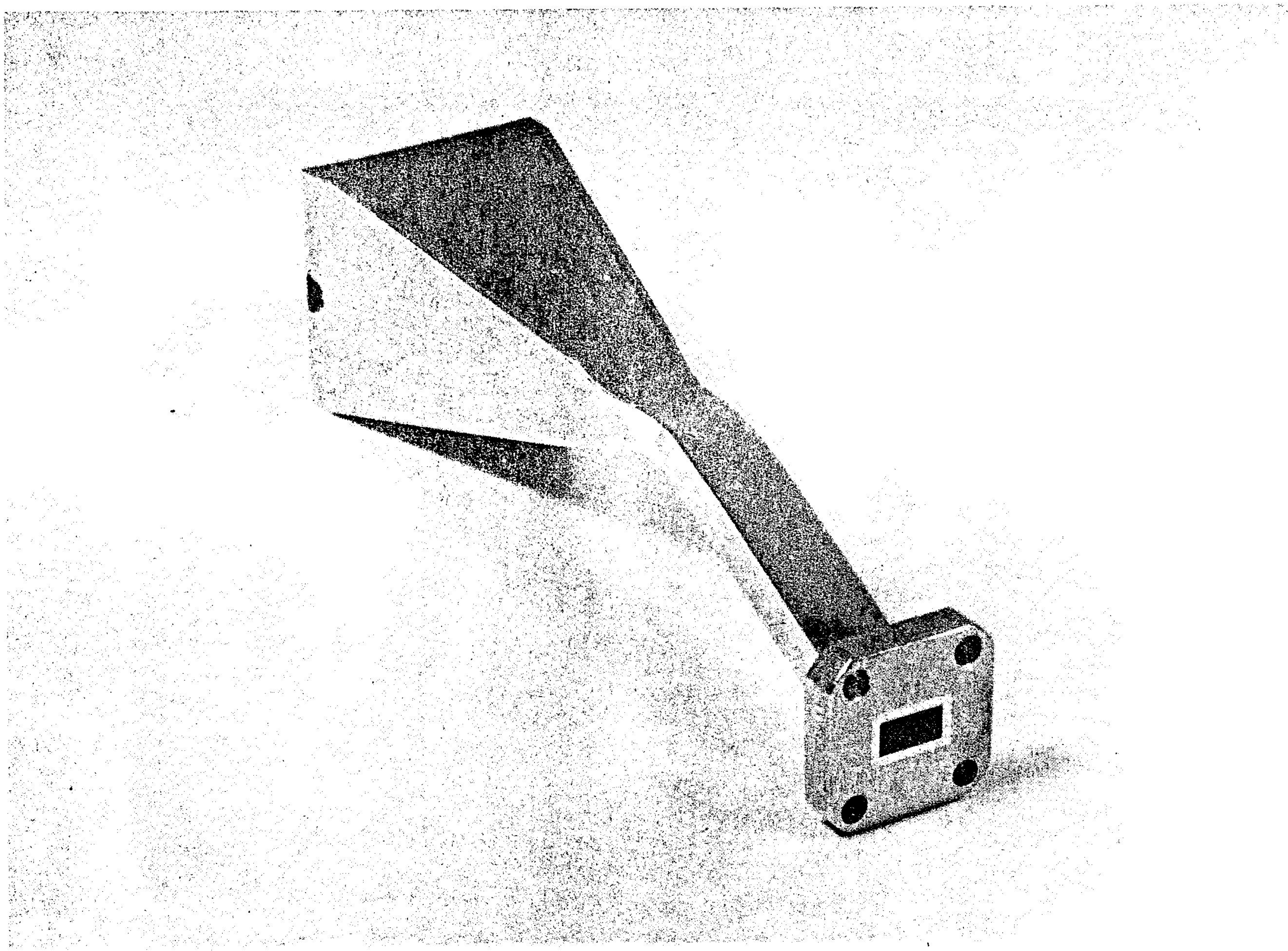


Figure 23 - Single Compensated Horn and Feed Waveguide
with Offset Bends in Both Planes.

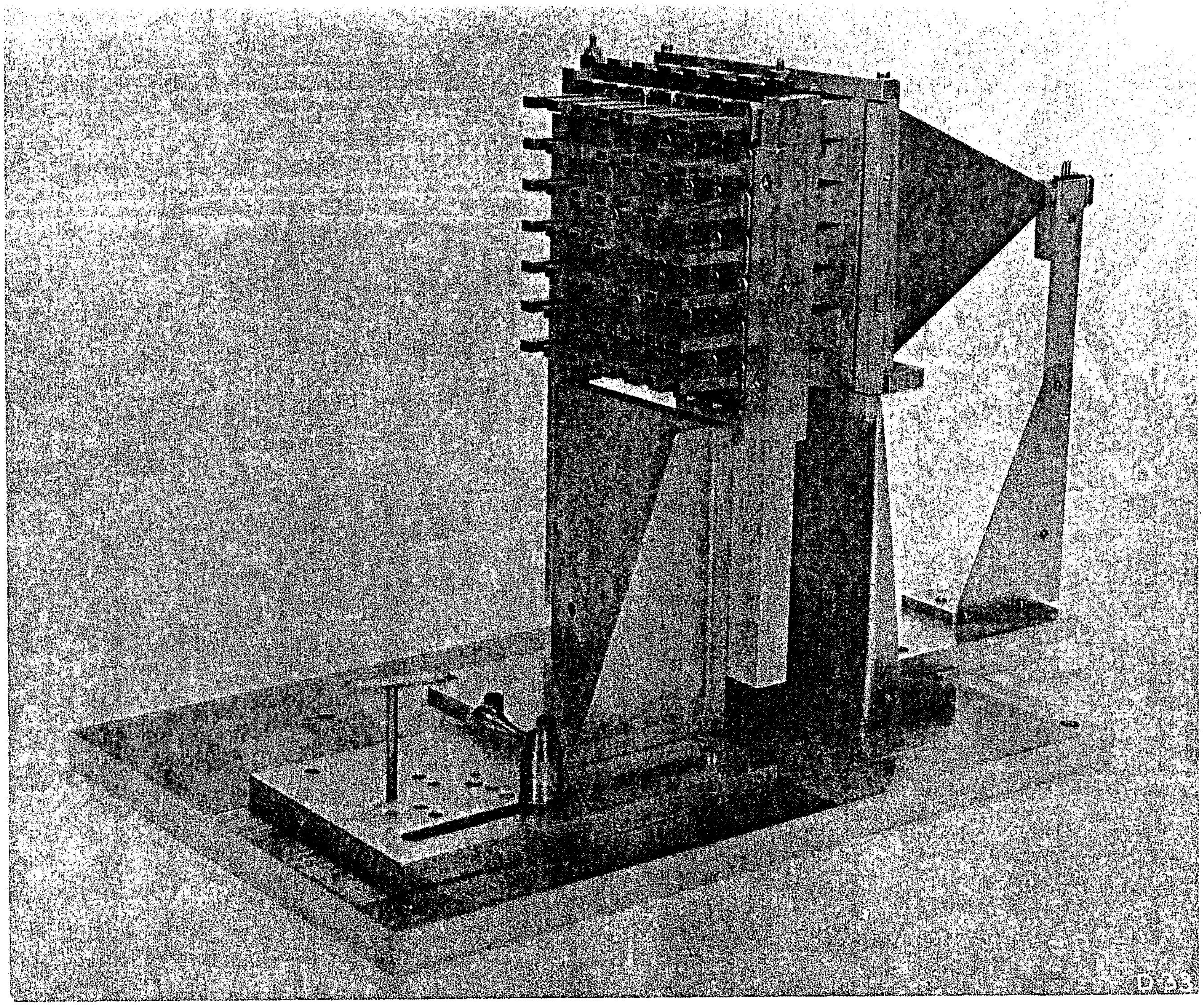


Figure 24 - Test Fixture and Optical Feed System

illuminating horn in many different modes. The slightest distortion of the physical dimensions of the horn (such as that which could be induced by fairly light finger pressure on the broad walls) had a marked effect on the energy received at any horn. Quantitatively, this effect could produce in excess of 10 dB variation at some output ports. As will be indicated in later figures, a similar result could be obtained by slight variations in frequency. None of these variations could be explained by any simple models. The only inference available was that an extremely complex field structure was being set up within the large sectoral horn as a result of reflections from the smaller horns at the aperture.

The next step taken was one of trying to reduce the reflection problem by establishing a field pattern within a section of oversized waveguide which would be more compatible with the terminating structure. A transition section was fabricated for insertion between the illuminating horn and the collecting horns; this section was of simple rectangular cross section. Within this section, dielectric slabs were inserted using a low dielectric constant (≈ 1.02) foam material for support. The slabs were designed so that each represented a transition to a type of H-guide wherein most of the energy would be propagating in or near the higher dielectric constant material in the hybrid H-guide mode. It was hoped that this would have the effect of more closely approximating the amplitude configuration which the boundary conditions at the seven horns would require and thus minimize the reflection problem. The structure is shown in Figure 25.

If the reflections were at all reduced by this effort, the effect was masked completely by the increased Q of the longer structure. The results of these various attempts (at one output port) are summarized in Figures 26 and 27. The first compares the initial problem in two polarizations; the second shows the result of trying to

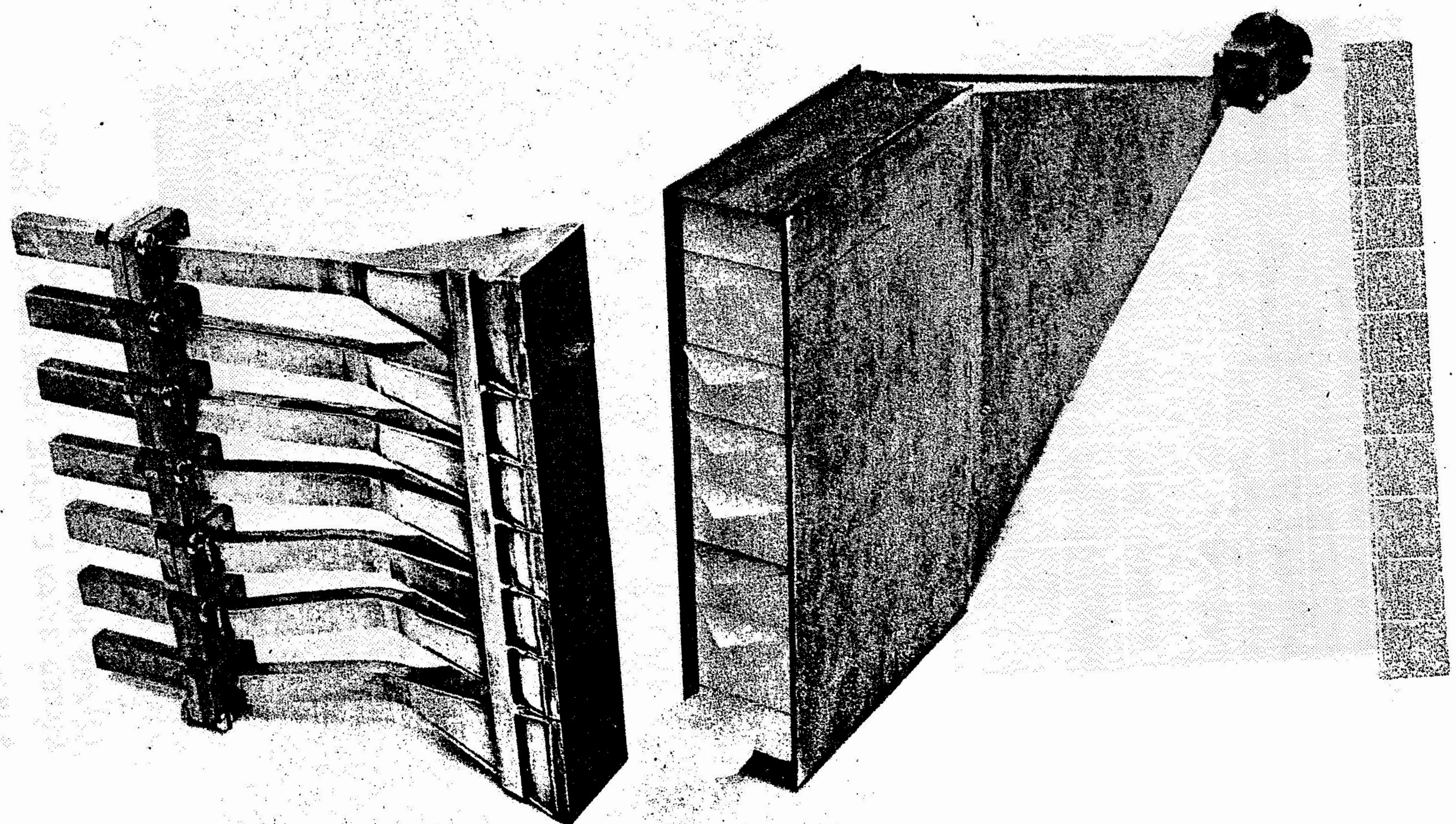
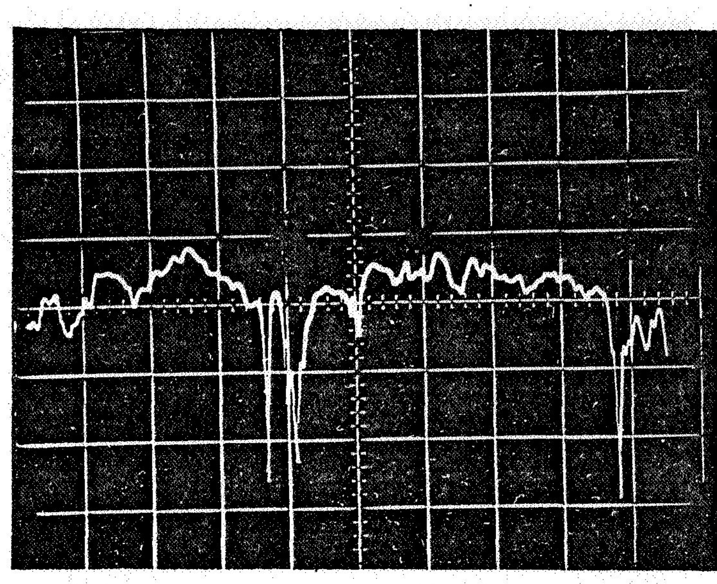
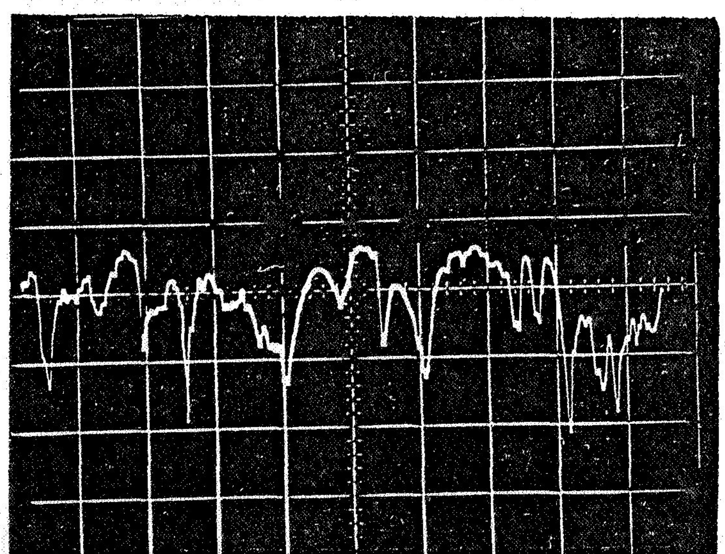


Figure 25 - Sectoral Horn and Transition Section with Dielectric Slabs



CENTER HORN OF SEVEN UNIFORM HORNS ILLUMINATED BY AN H-PLANE FLARE SECTORAL HORN

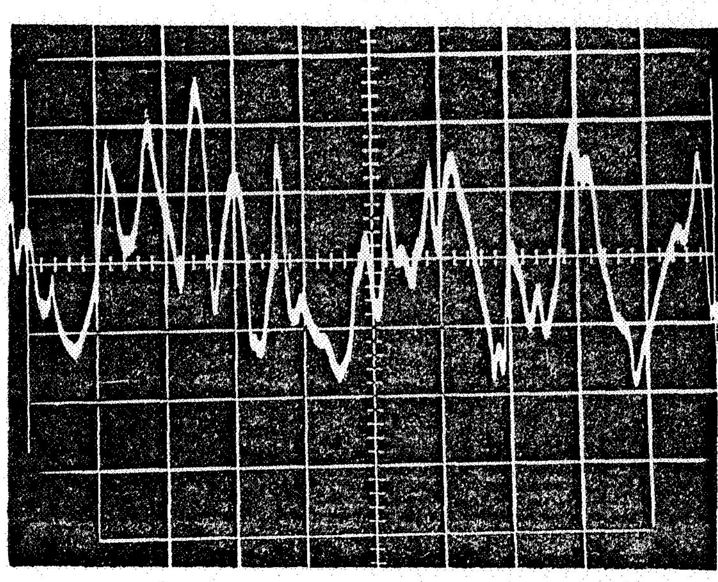
5 dB/cm 34.5 GHz TO 35.5 GHz



CENTER HORN OF SEVEN UNIFORM HORNS ILLUMINATED BY AN E-PLANE FLARE SECTORAL HORN

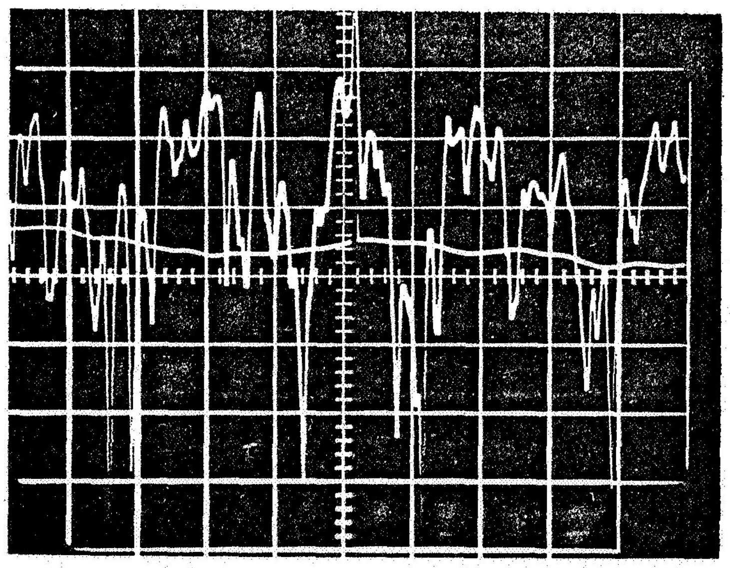
5 dB/cm 34.5 GHz TO 35.5 GHz

FIG. 26 - COMPARISON OF RESPONSES USING DIFFERENT POLARIZATIONS FOR THE SECTORAL FEED HORNS



CENTER HORN OF SEVEN NON-UNIFORM HORNS ILLUMINATED BY AN H-PLANE FLARE SECTORAL HORN

2 dB/cm 34.5 GHz TO 35.5 GHz



CENTER HORN OF SEVEN NON-UNIFORM HORNS ILLUMINATED BY AN H-PLANE FLARE SECTORAL HORN WITH DIELECTRIC SLAB FOR AMPLITUDE TAILORING

2 dB/cm 34.5GHz TO 35.5 GHz

FIG. 27 - COMPARISON OF RESPONSES USING SECTORAL HORN AND NONUNIFORM COLLECTING HORN

compensate with the H-guide approach. The frequency scan is 1 GHz also and the final trace shows the detector response, in the absense of the horn assembly, superimposed.

Although time was running out, we were reluctant to abandon this type of approach inasmuch as it seems much more attractive for scaling to 94 GHz than does the series feed. Hence, one final effort was made wherein one sectoral horn was modified to accept a radial array of uniform horns. The axis of each collecting horn was then aligned with a ray to the phase center of the illuminating horn. This configuration is shown in Figure 28. The power output at the center horn is then shown in Figure 29. It can be seen that this modification reduced the power variations from 10 dB to the order of 2 dB. Further, the variation from horn to horn was considerably less in this configuration than in any other. Measurements showed only a 2 dB taper across the array which, in this polarization, should be uniform. In addition, the system was 95% efficient.

With this final bit of modest success, we believe that a configuration along these lines could be developed to serve as well or perhaps better than the slot array as a multiple power dividing system. The concept is illustrated in Figure 30. The sectoral horn shown would then feed seven more identical structures which would be attached to the twists. The result is a greatly reduced 7 x 7 matrix of waveguides which could flare to the .771" spacing for the phase shifters and output array.

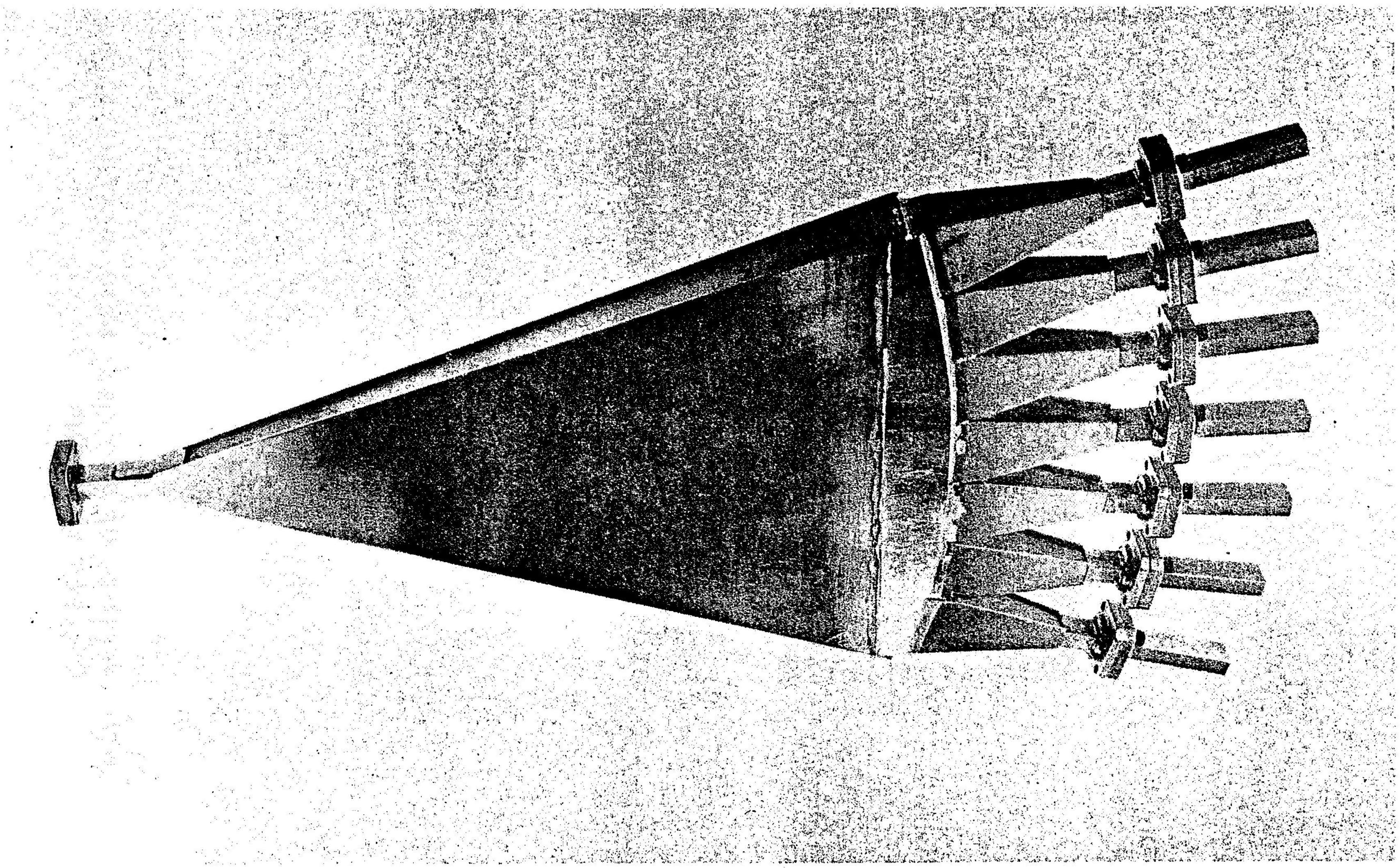
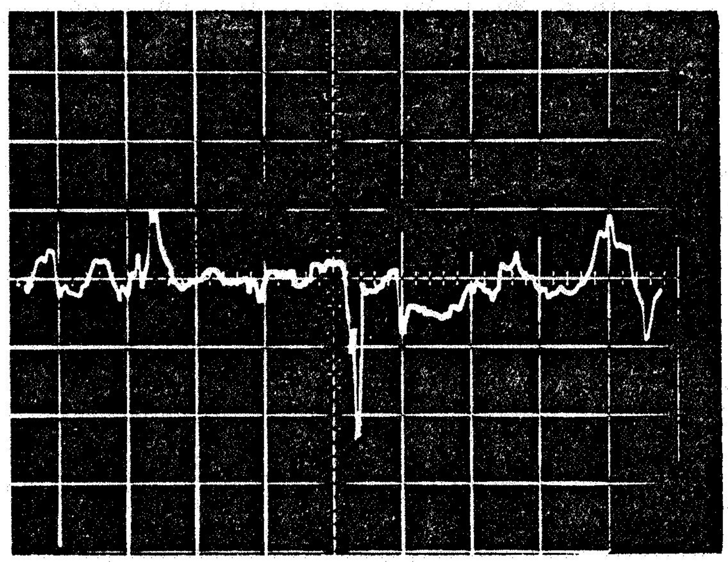


Figure 28 - Sectoral Feed and Radially Mounted Collecting Horns



CENTER HORN OF SEVEN UNIFORM HORNS MOUNTED SO THEIR APERTURES ARE PERPENDICULAR TO THE IMPINGING AXIAL RAY FROM THE E-PLANE SECTORAL HORN

2 dB/cm 34.5 GHz TO 35.5 GHz

FIG. 29-RESULTS OF SECTORAL HORN FEEDING RADIAL COLLECTING HORNS

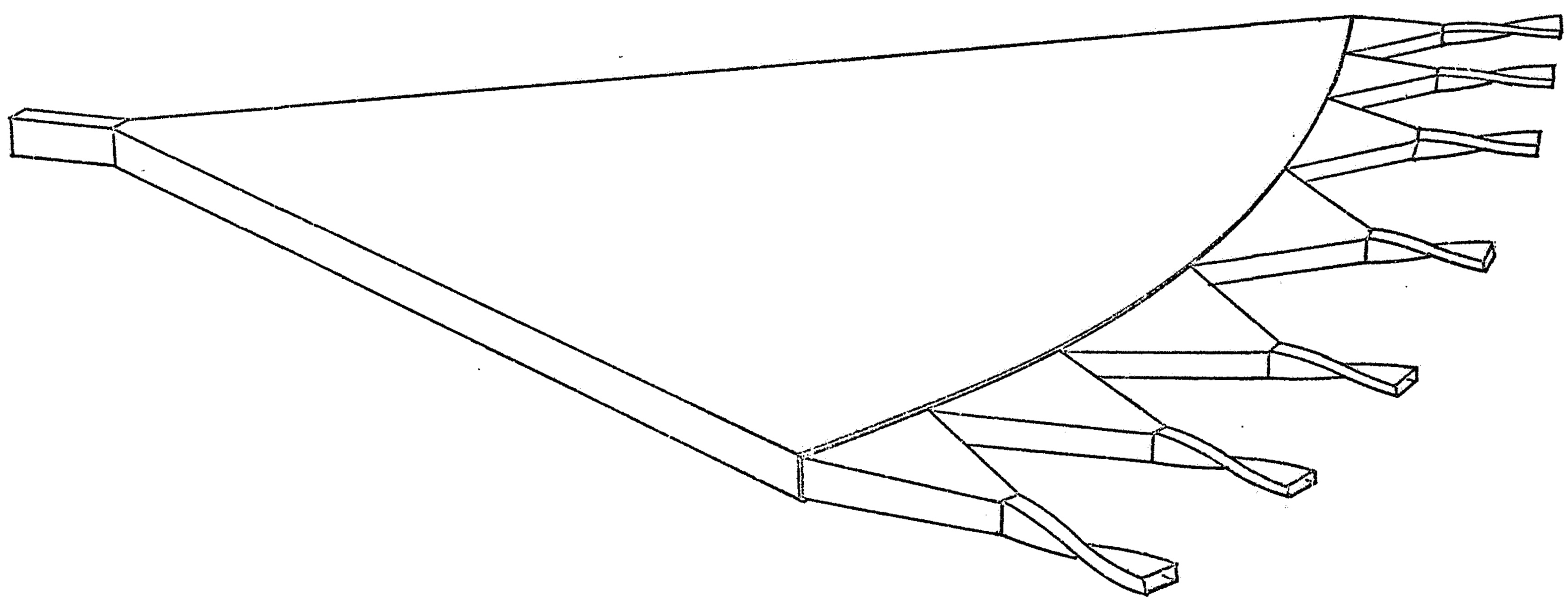


FIG. 30 - CONCEPT OF SECTORAL HORN OPTICAL ILLUMINATION SYSTEM

7. SYSTEM TEST RESULTS

Although the quasi-optical feed was not sufficiently developed to be tested with the phase shifter and output array, the series coupler system was tested and the method and results are reported below.

7.1 Method of Testing

In order to measure the radiation pattern of the series fed system it was first necessary to establish a uniform phase across the row of waveguides. This was done by first measuring the relative phase at each port with the phase shifter attached but in the "0" state. Then by pulsing in the necessary delay at each port a uniform phase was established for the transmit mode. However, since the phase shifters were non-reciprocal it was necessary to pulse in the "complementary word" to each phase shifter to enable it to receive. The feed arm and phase shifters were then bolted to a holding fixture and the uniform output horns attached to the phase shifters. (The normal arrangement of horns on the seven port coupler is such that they are in the E-plane polarization i.e. the amplitude distribution across each horn seen along the row is uniform.) The system was then mounted on ADTEC's rooftop antenna pattern range and the radiation pattern measured over $\pm 30^{\circ}$ azimuth. Figure 31 illustrates the system mounted on the mast and Figure 32 illustrates the pattern range and system under test. In order to test the array in the H-plane polarization step twists were inserted between each phase shifter and horn.

7.1.1 E-Plane Radiation Patterns

Figure 33 is a plot of the computer results for a 35 GHz Boresight pattern. Figures 34 to 41 are the measured patterns. The figures are cropped so as to show the maximum sidelobe within the angular range of interest, and the magnitude of this sidelobe is given for reference.

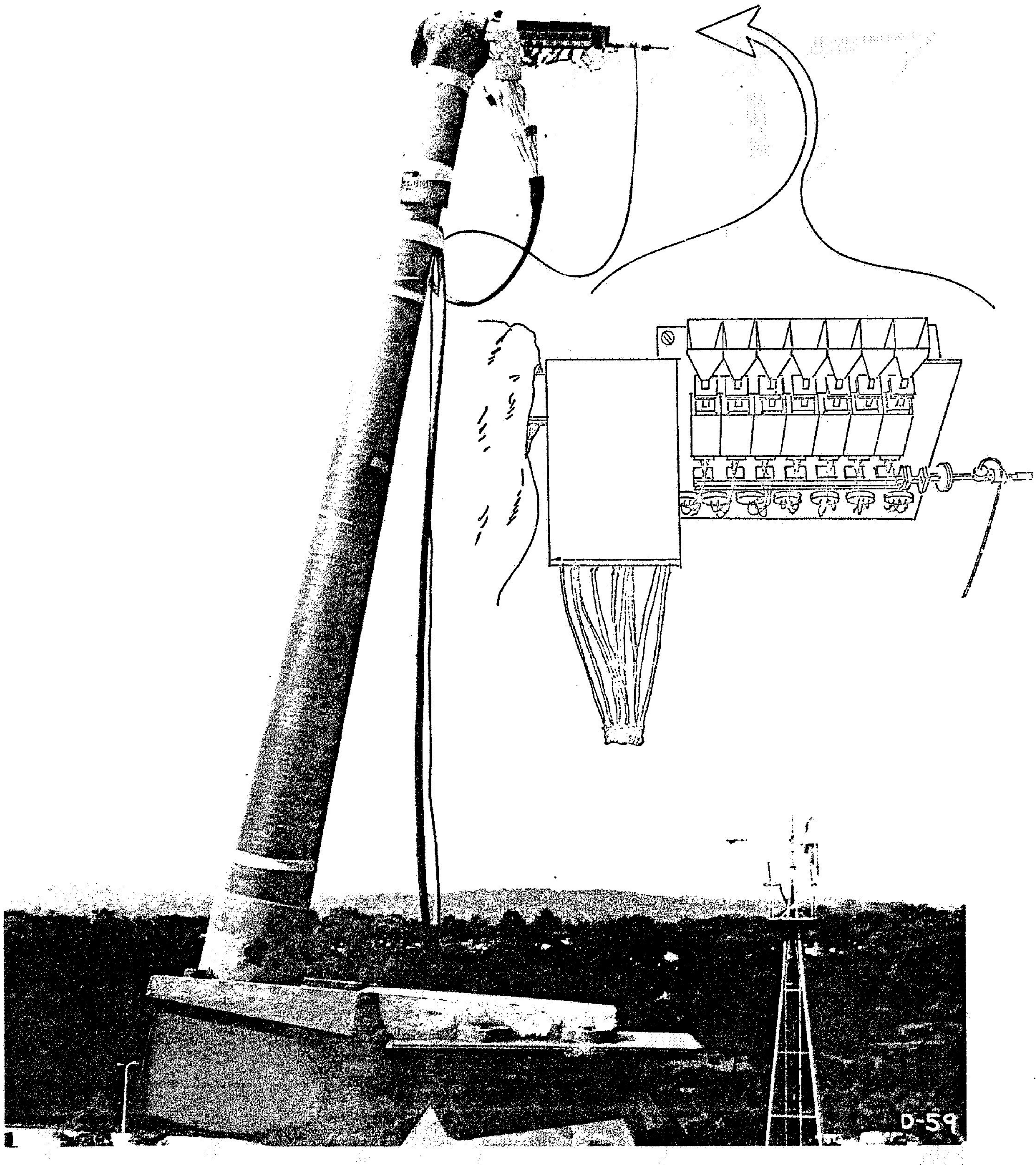


Figure 31 - Series Fed Antenna System Mounted on Mast

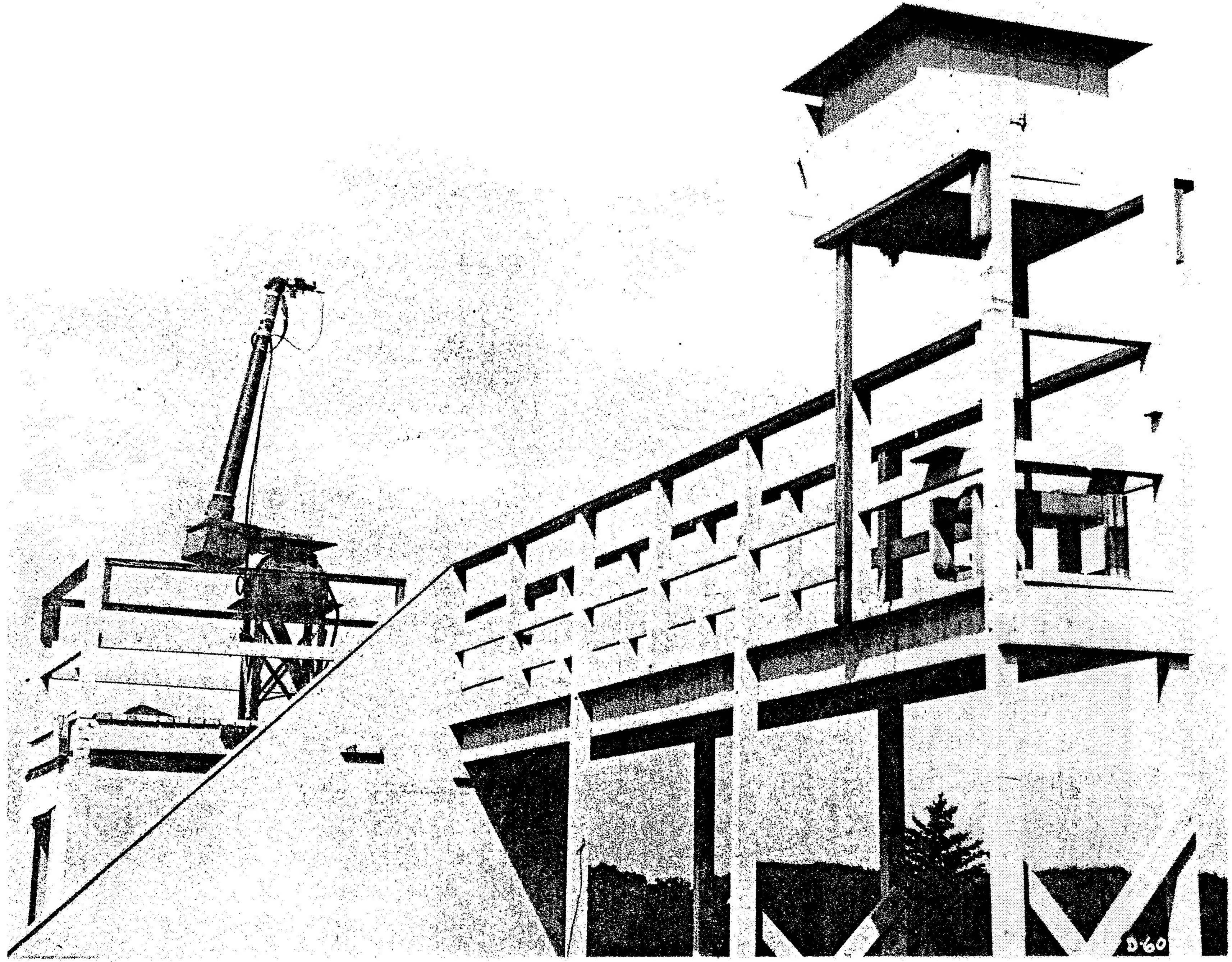


Figure 32 - Antenna Pattern Range and System Under Test

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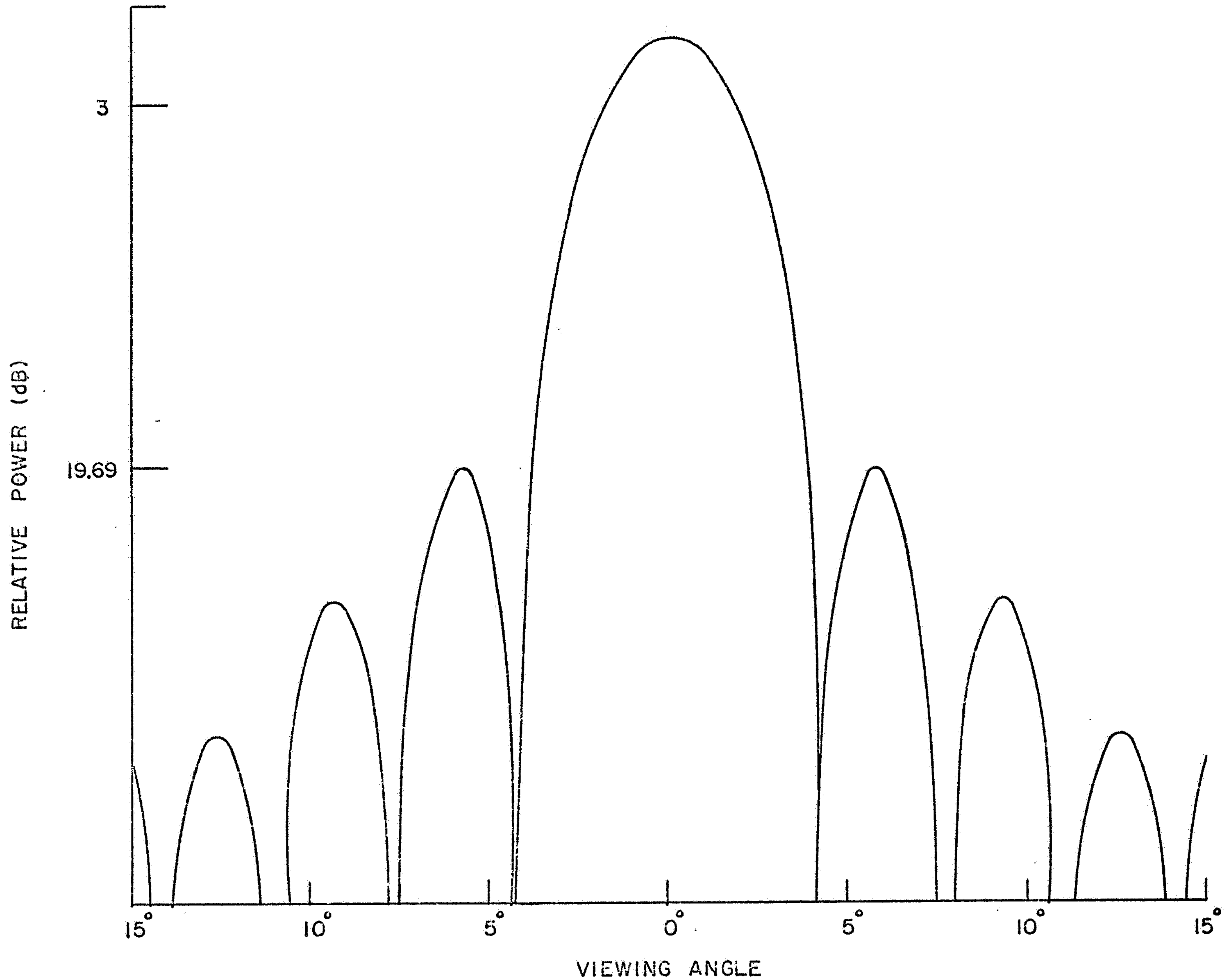

FIG. 33-THEORETICAL PATTERN LINEAR ARRAY E-PLANE POLARIZATION 35.0 GHz BORESIGHT

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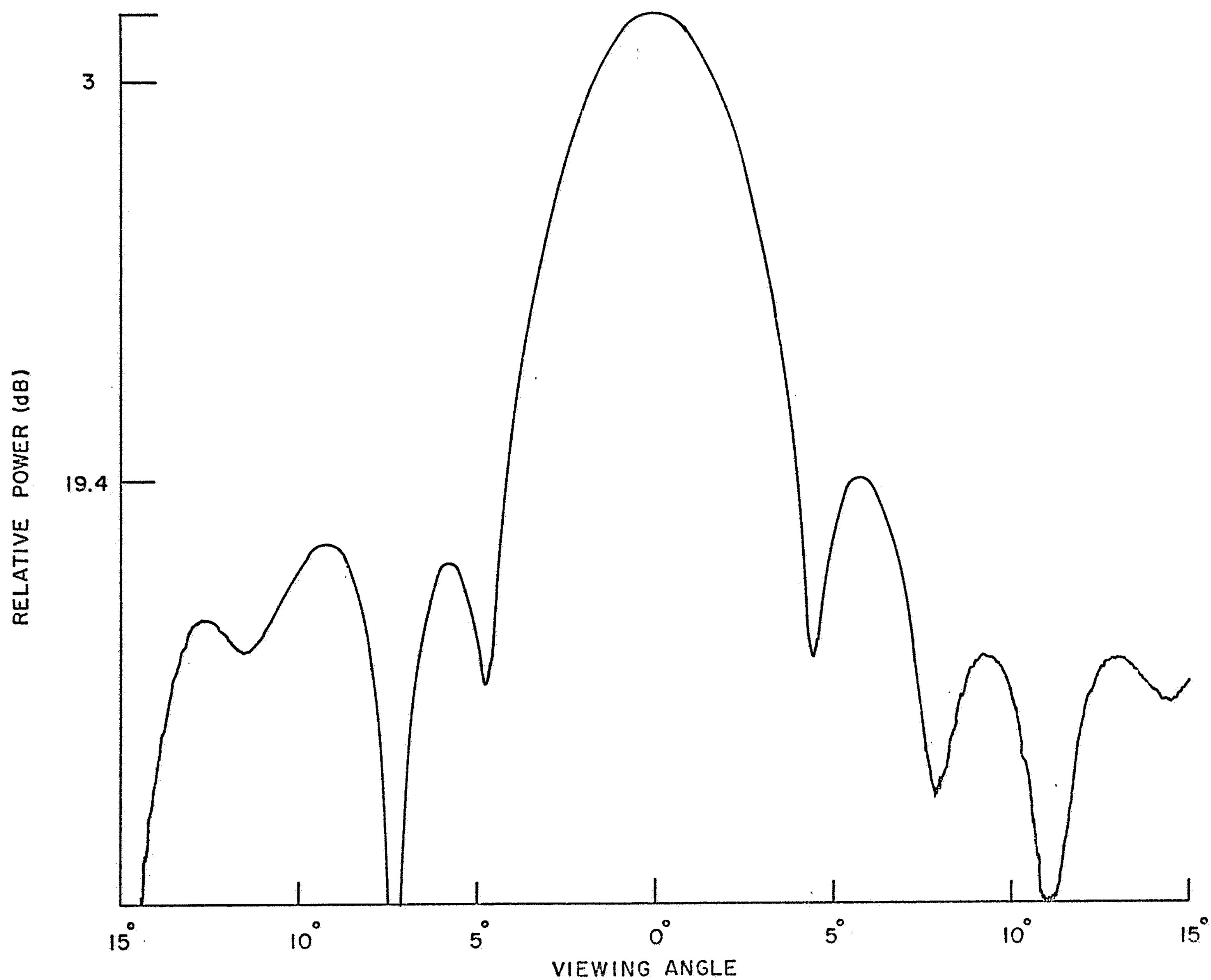

FIG. 34-LINEAR ARRAY E-PLANE POLARIZATION 35 GHz BORESIGHT

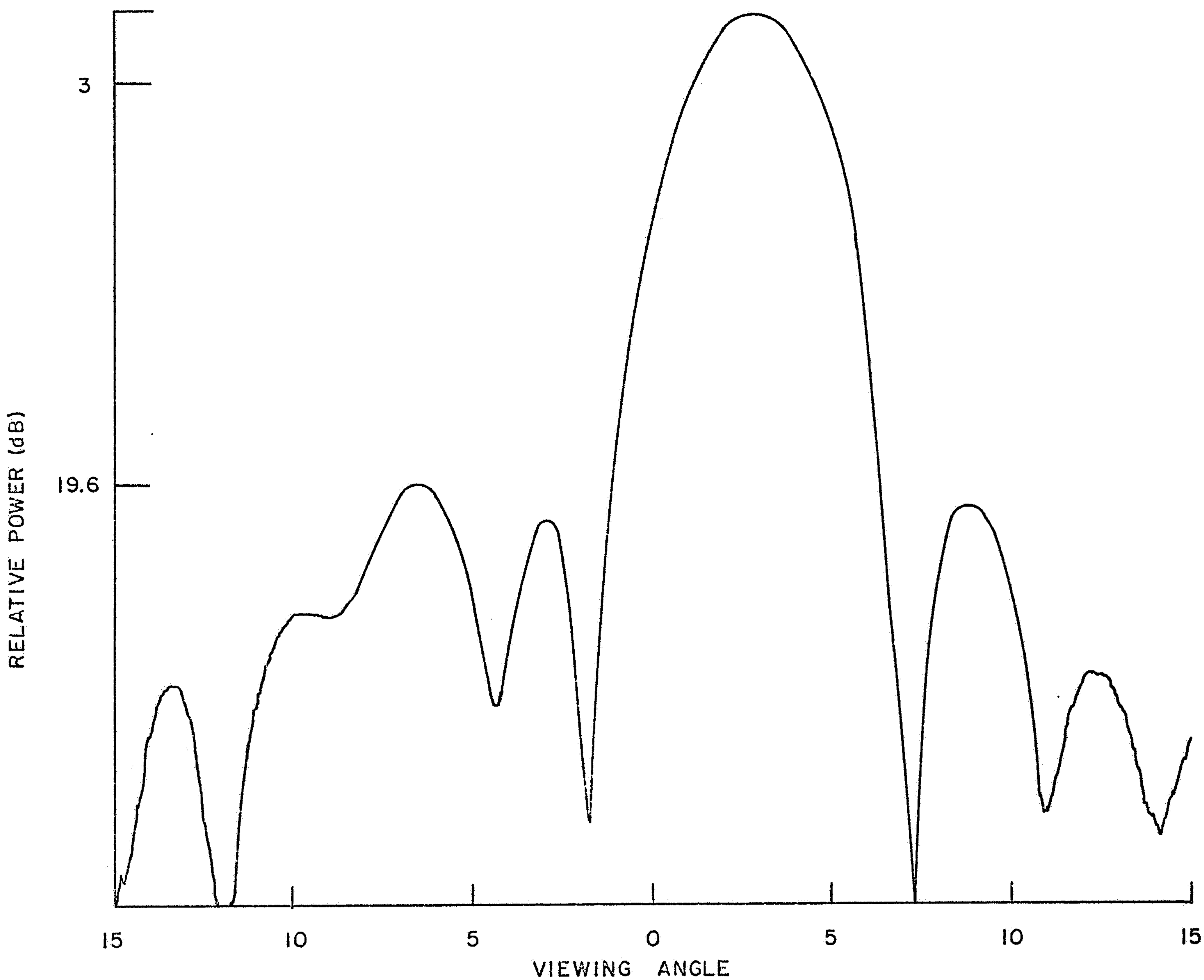

FIG. 35-LINEAR ARRAY E-PLANE POLARIZATION 35 GHz STEERED 3.2°

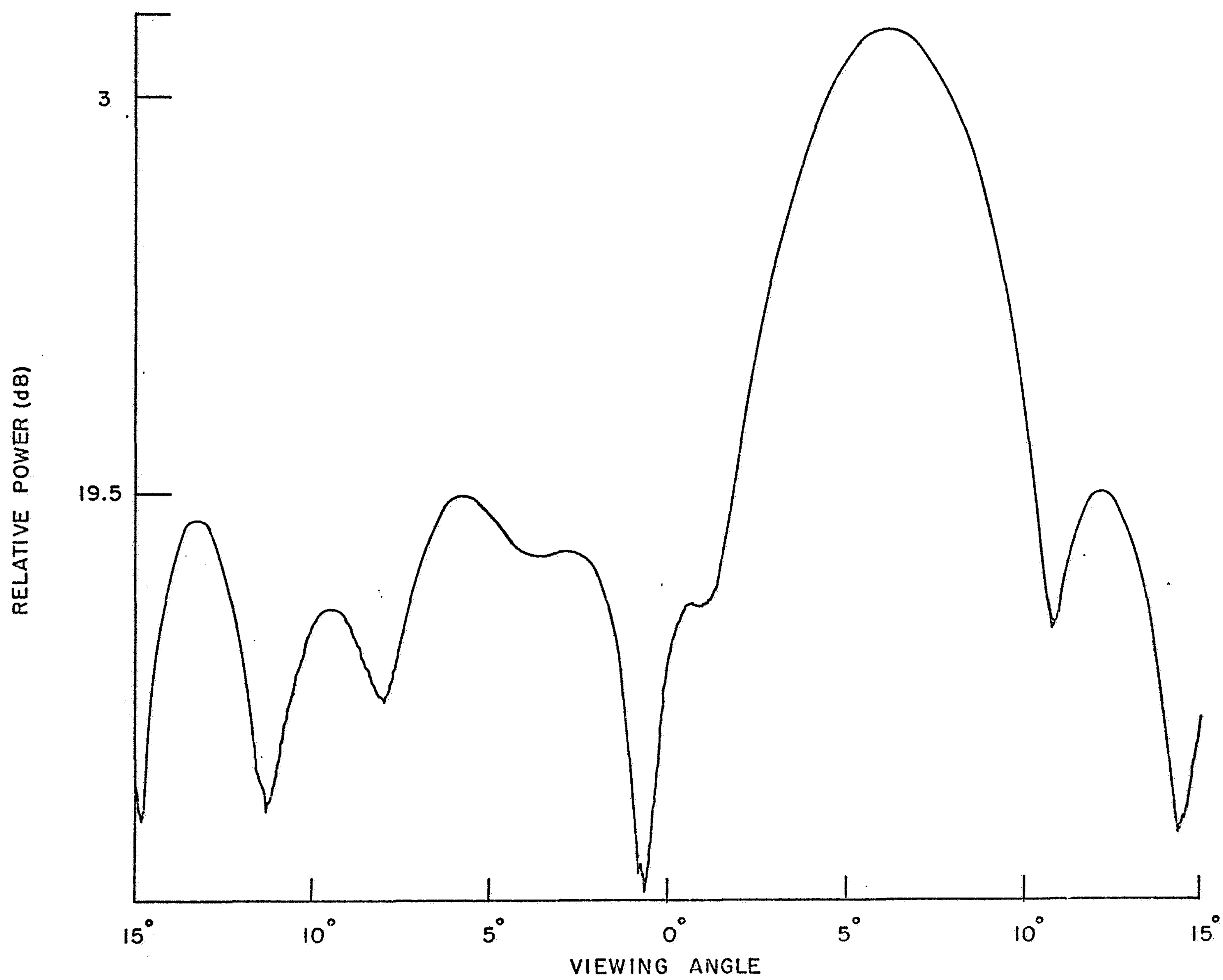

FIG.36-LINEAR ARRAY E-PLANE POLARIZATION 35 GHz STEERED 6.4°

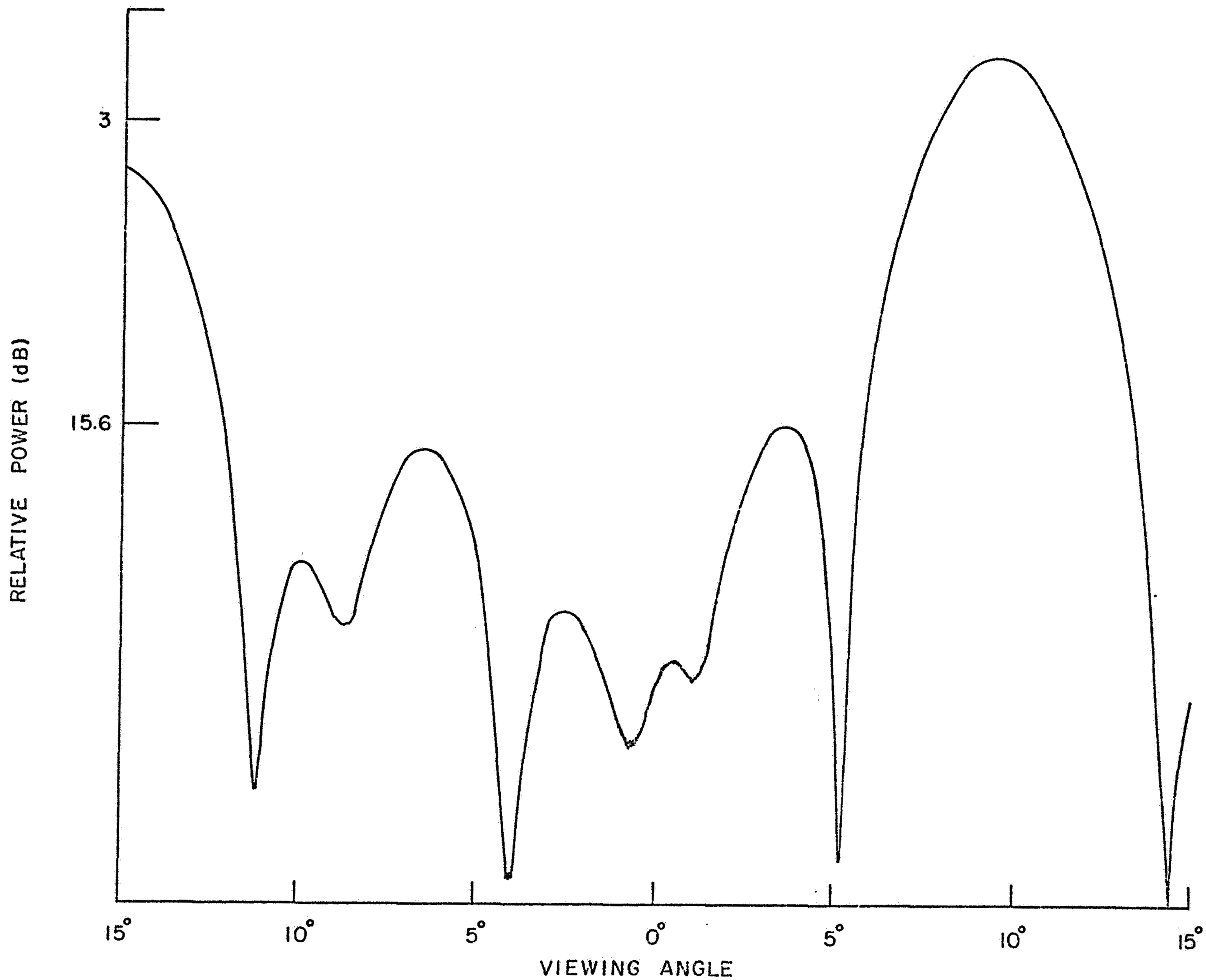

FIG. 37-LINEAR ARRAY E-PLANE POLARIZATION 35GHz STEERED 9.6°

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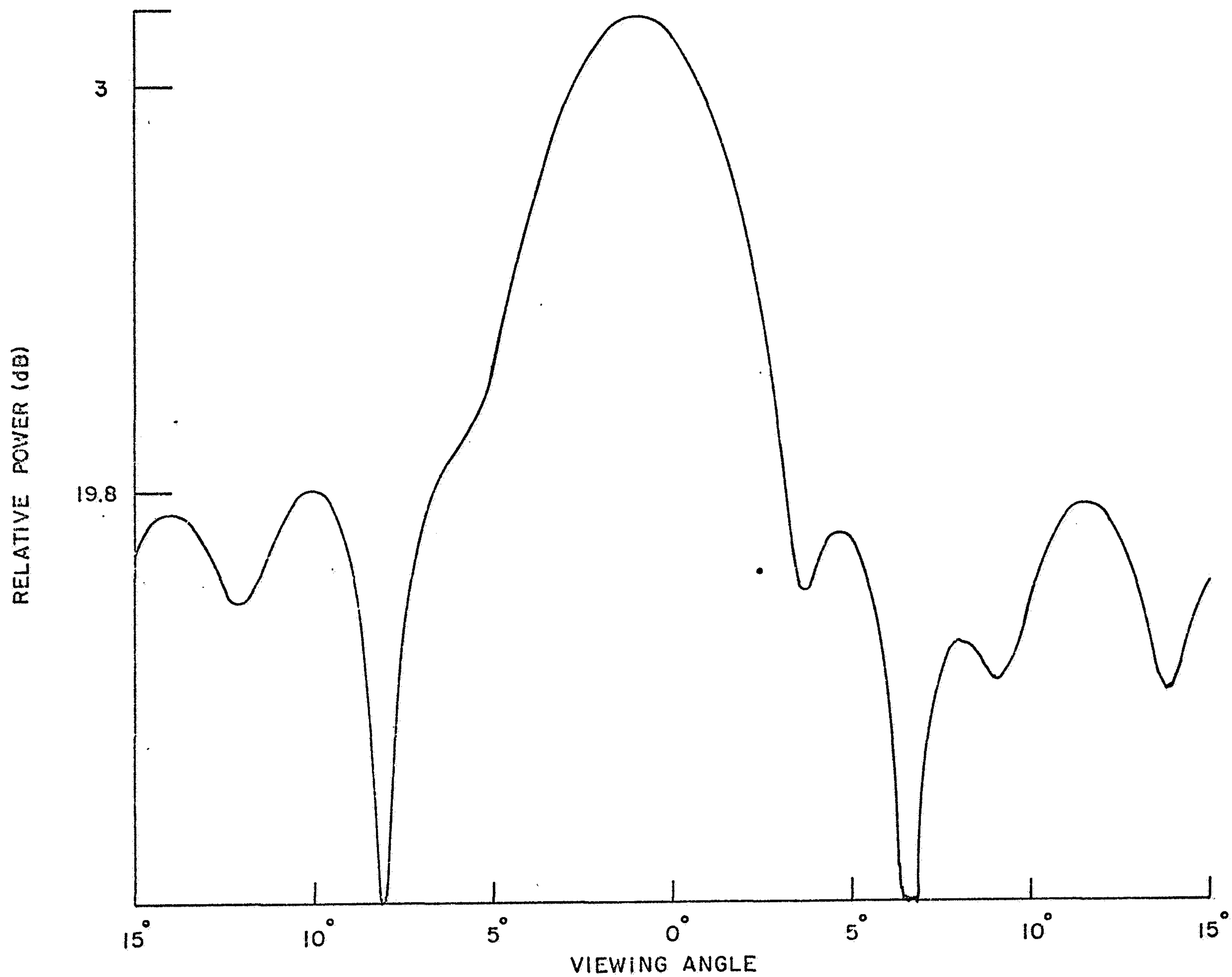

FIG.38-LINEAR ARRAY E-PLANE POLARIZATION 36.4 GHz BORESIGHT

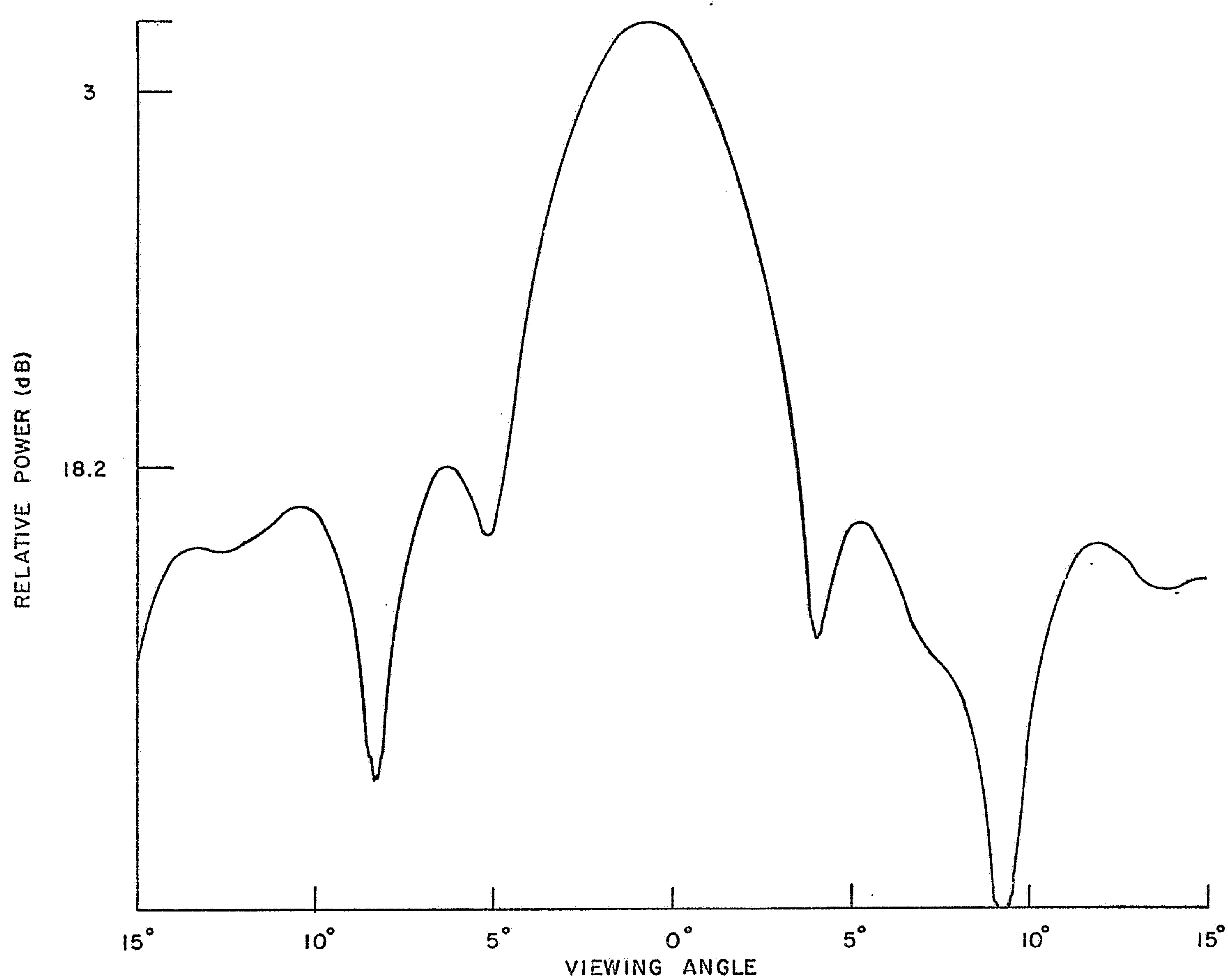

FIG. 39-LINEAR ARRAY E-PLANE POLARIZATION 34.8GHz BORESIGHT

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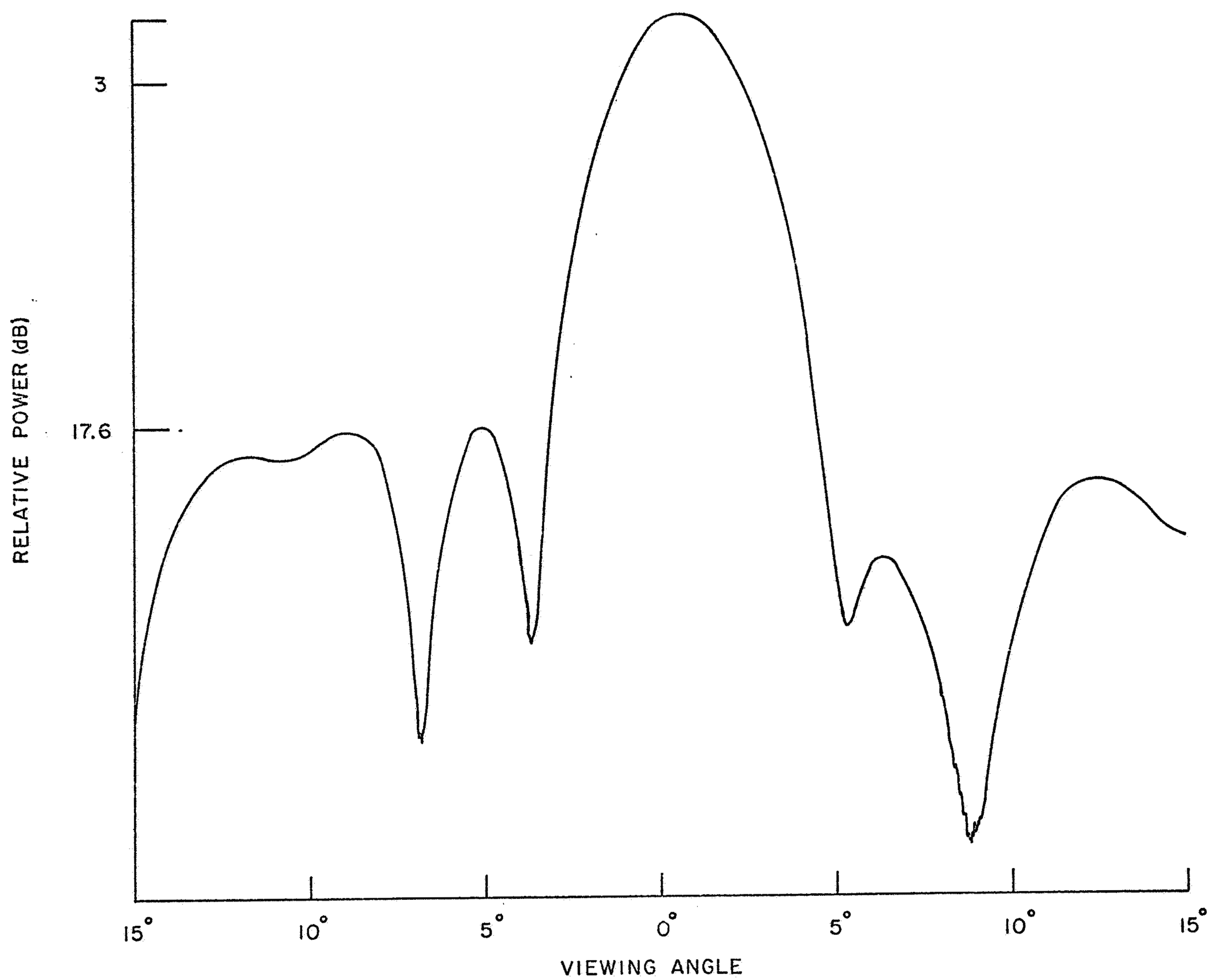

FIG.40-LINEAR ARRAY E-PLANE POLARIZATION 35.2 GHz BORESIGHT

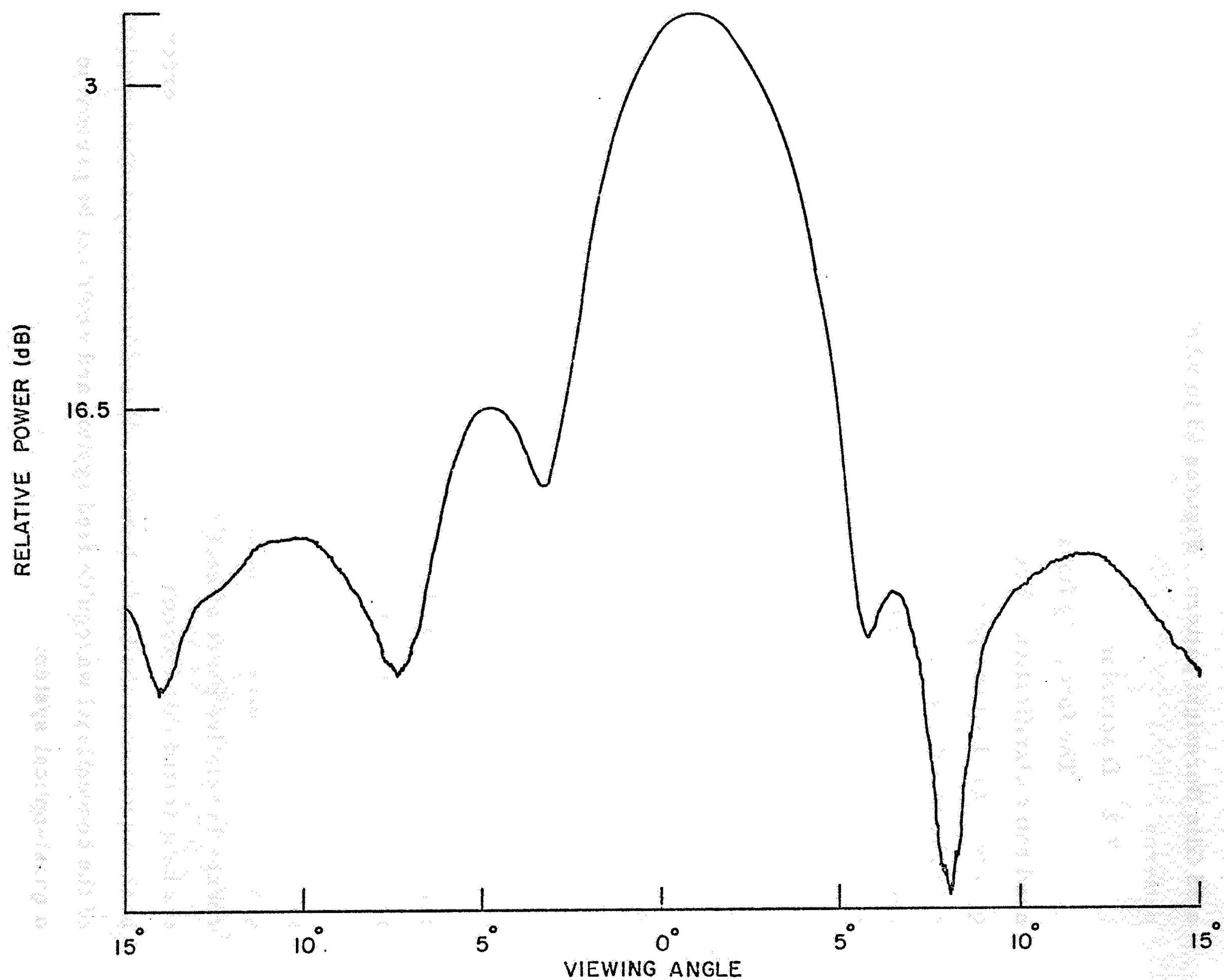

FIG. 41 - LINEAR ARRAY E-PLANE POLARIZATION 35.4 GHz BORESIGHT

7.1.2 H-Plane Radiation Patterns

Figure 42 is a plot of the computer results for a 35 GHz Boresight pattern. Figures 43 to 50 are the measured patterns.

7.2 Discussion

The foregoing figures although mostly self explanatory need some clarification. The boresight pattern at the design frequency in the E-plane polarization is quite acceptable since the highest sidelobe is only .2 dB above the theoretical value. The steered patterns are also very close to the theoretical. The beamwidth is approximately 3.8° as compared to the 3.6° for the ideal and the grating lobe (not shown in the figure) appears at 26° and well removed from the region of interest. The latter remained outside the field of interest even in the extreme steering condition.

However, problems arise when the system is moved from the design frequency. Nearly random variations in the insertion loss of the ferrite phase shifters cause a modification in the illumination function which raises the sidelobe levels at frequencies other than 35 GHz. The ferrite phase shifters, then, are the major band limiting element in the system. Since these are prototype models there is much room for improvement and it is expected that this will not be a problem in the future. It has been shown that the seven port coupler feed works well over the 34.5 to 35.5 GHz and does not contribute greatly to the degradation of the patterns. Another characteristic of changes in frequency is the small amount of steering that occurs due to a progressive change in relative phase at each port with changes in frequency. Over the band tested this frequency steering accounted for less than 1° error at the extremes of the band. However, this effect is only a characteristic of the conventional waveguide feed system and would not be present in a quasi-optical system.

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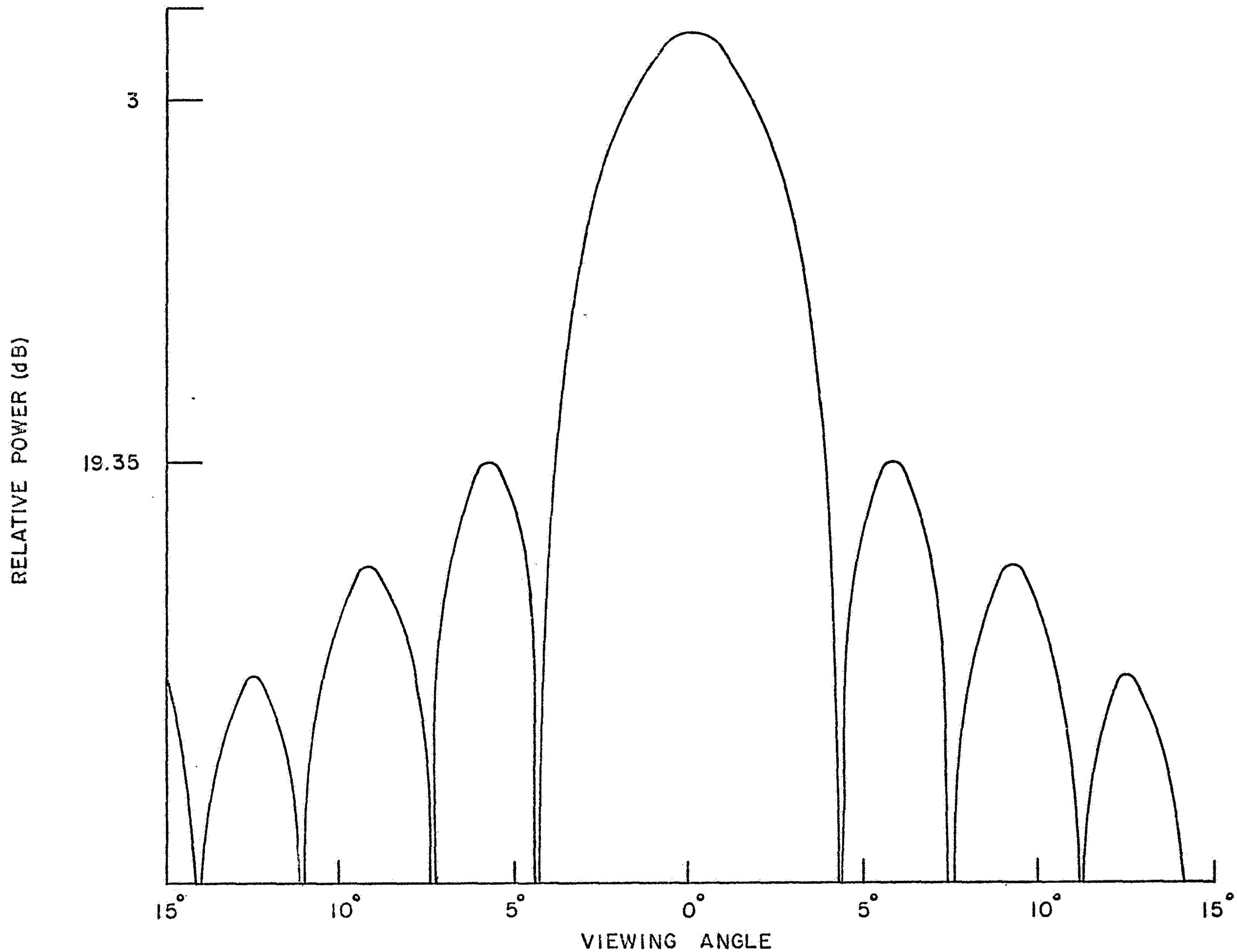

FIG. 42–THEORETICAL PATTERN LINEAR ARRAY H-PLANE POLARIZATION 35GHz BORESIGHT

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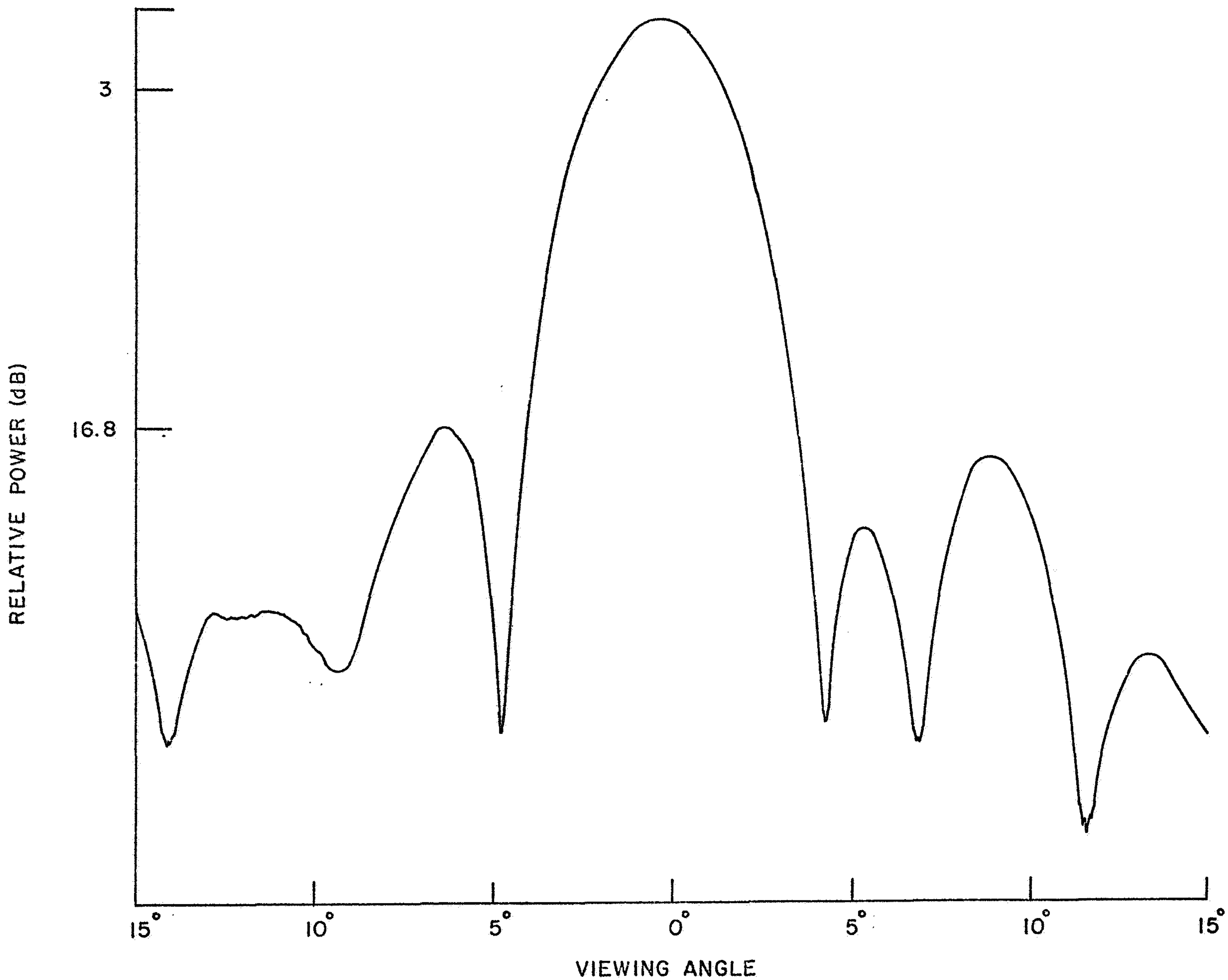

FIG. 43-LINEAR ARRAY H-PLANE POLARIZATION 35.02 GHz BORESIGHT

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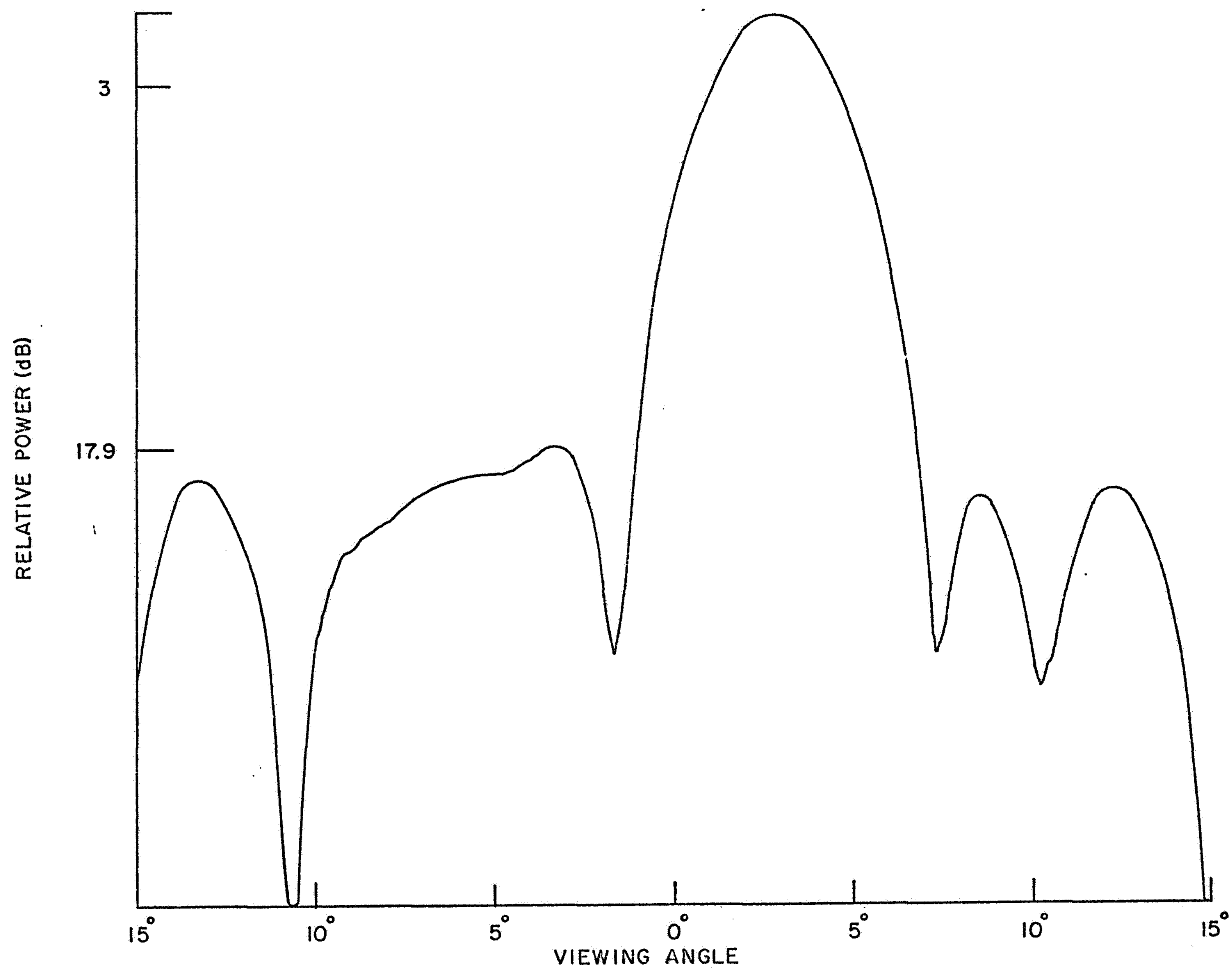

FIG. 44–LINEAR ARRAY H-PLANE POLARIZATION 35.02 STEERED 3.2°

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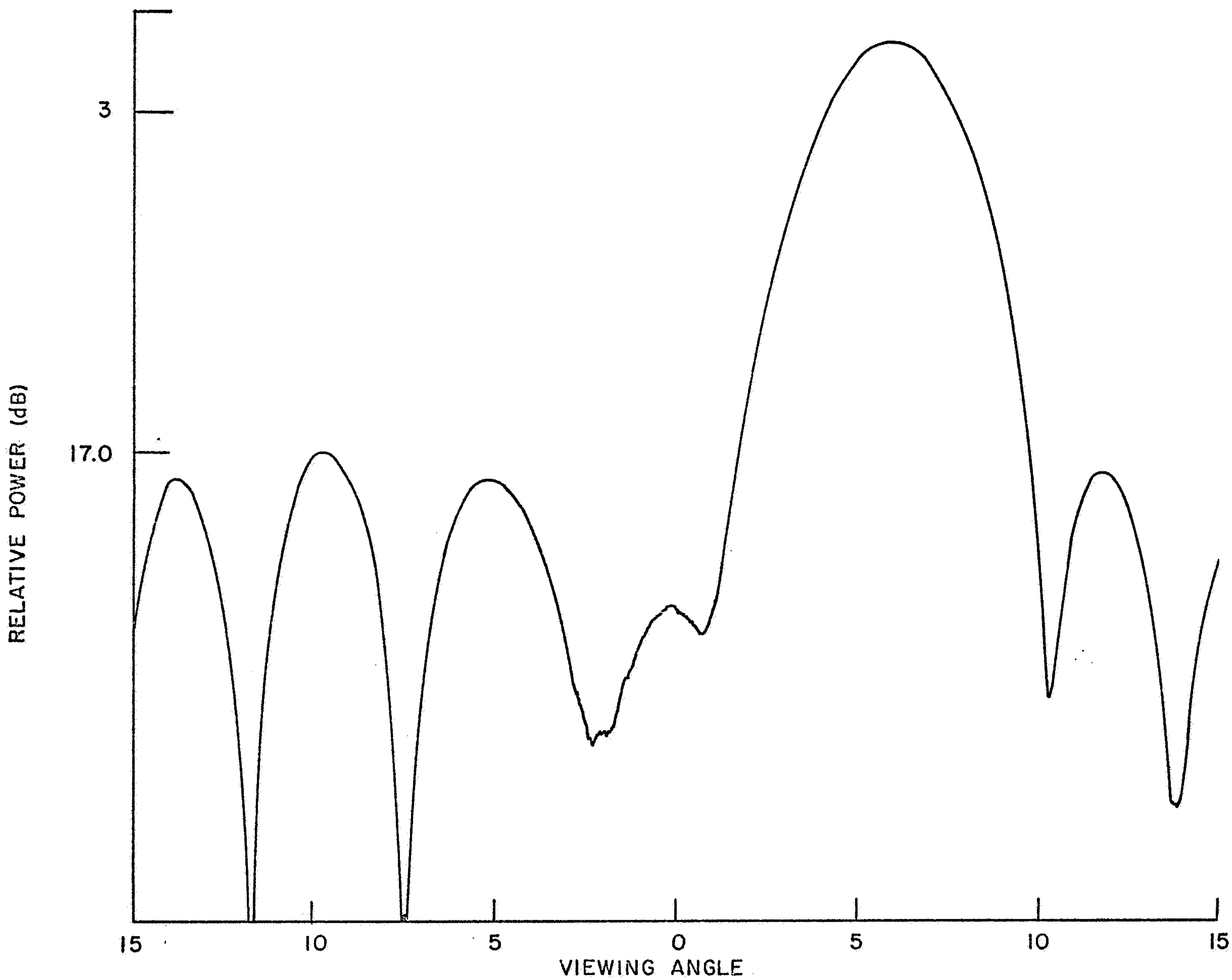

FIG. 45-LINEAR ARRAY H-PLANE POLARIZATION 35.02 GHz STEERED 6.4°

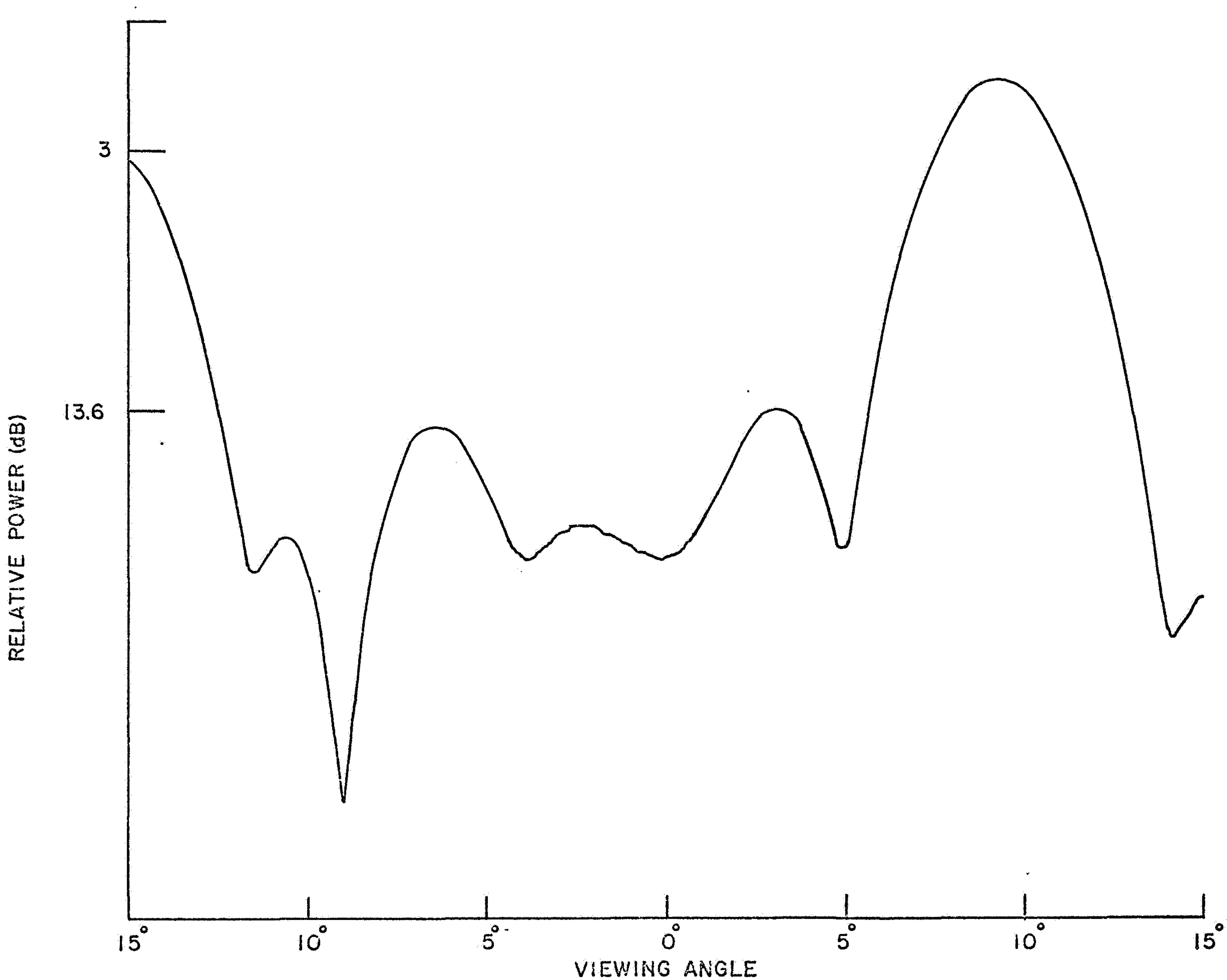

FIG.46-LINEAR ARRAY H-PLANE POLARIZATION 35.02 GHz STEERED 9.6°

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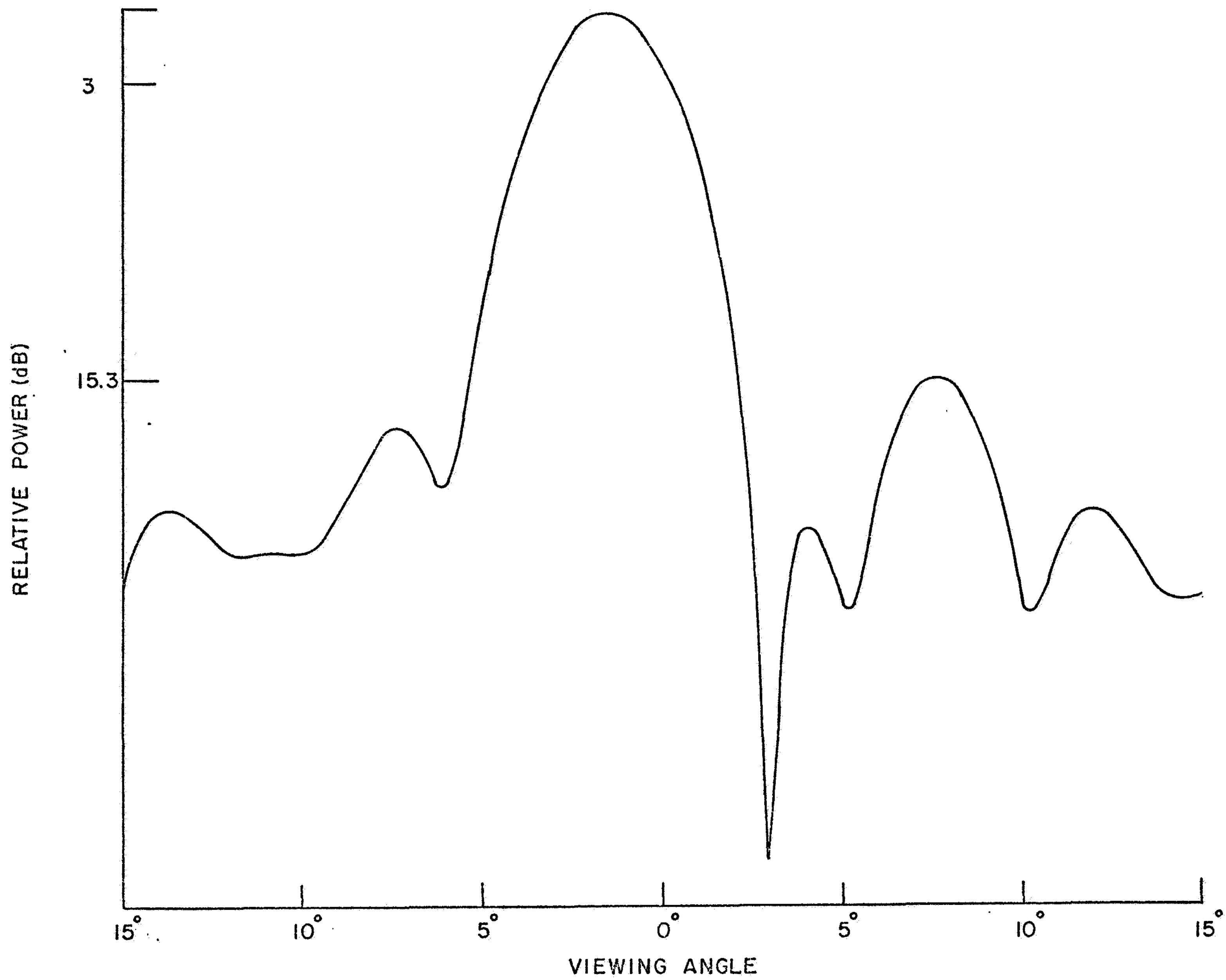

FIG.47-LINEAR ARRAY H-PLANE POLARIZATION 34.6 GHz BORESIGHT

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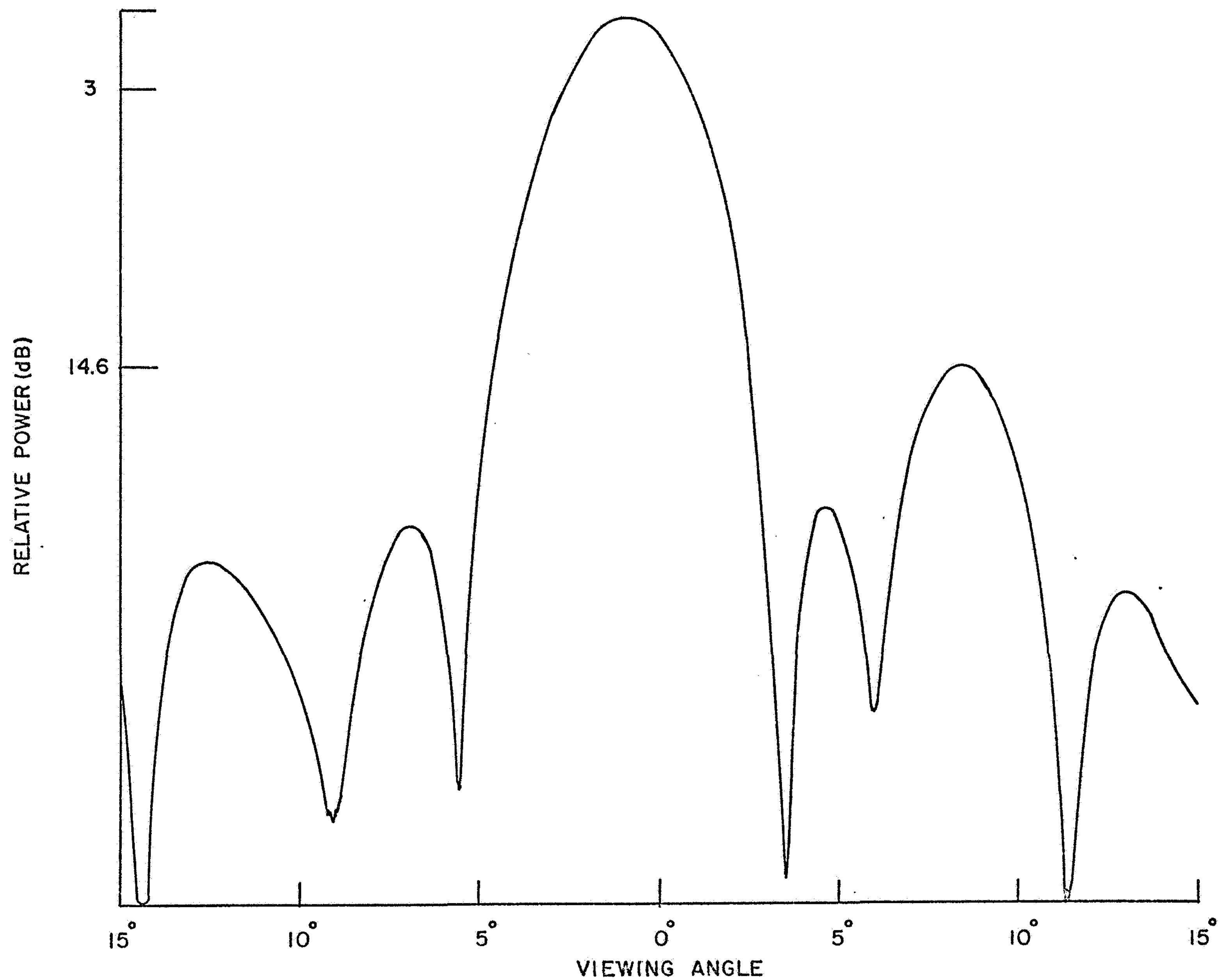

FIG. 48-LINEAR ARRY H-PLANE POLARIZATION 34.8 GHz BORESIGHT

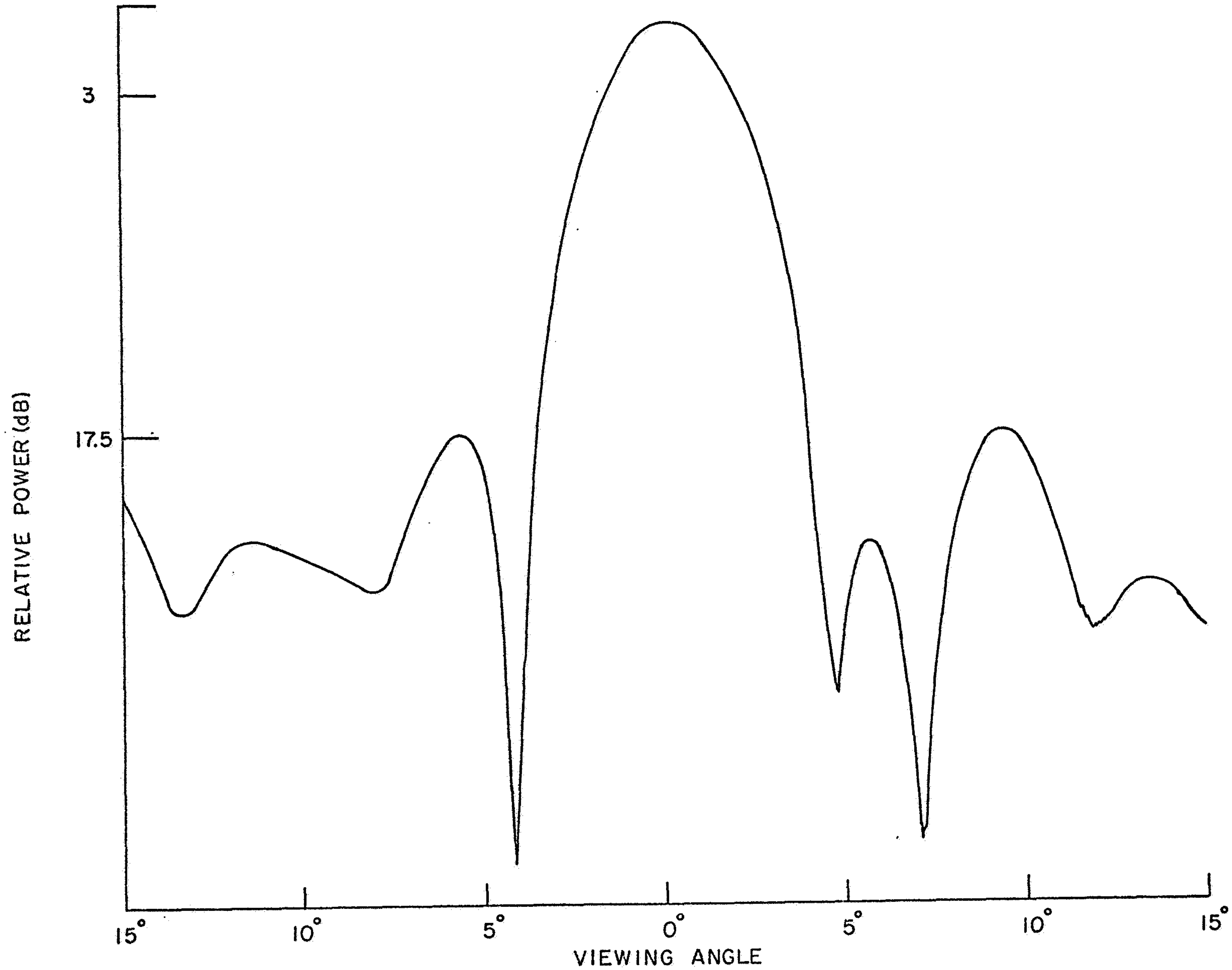

FIG. 49-LINEAR ARRAY H-PLANE POLARIZATION 35.2 GHz BORESIGHT

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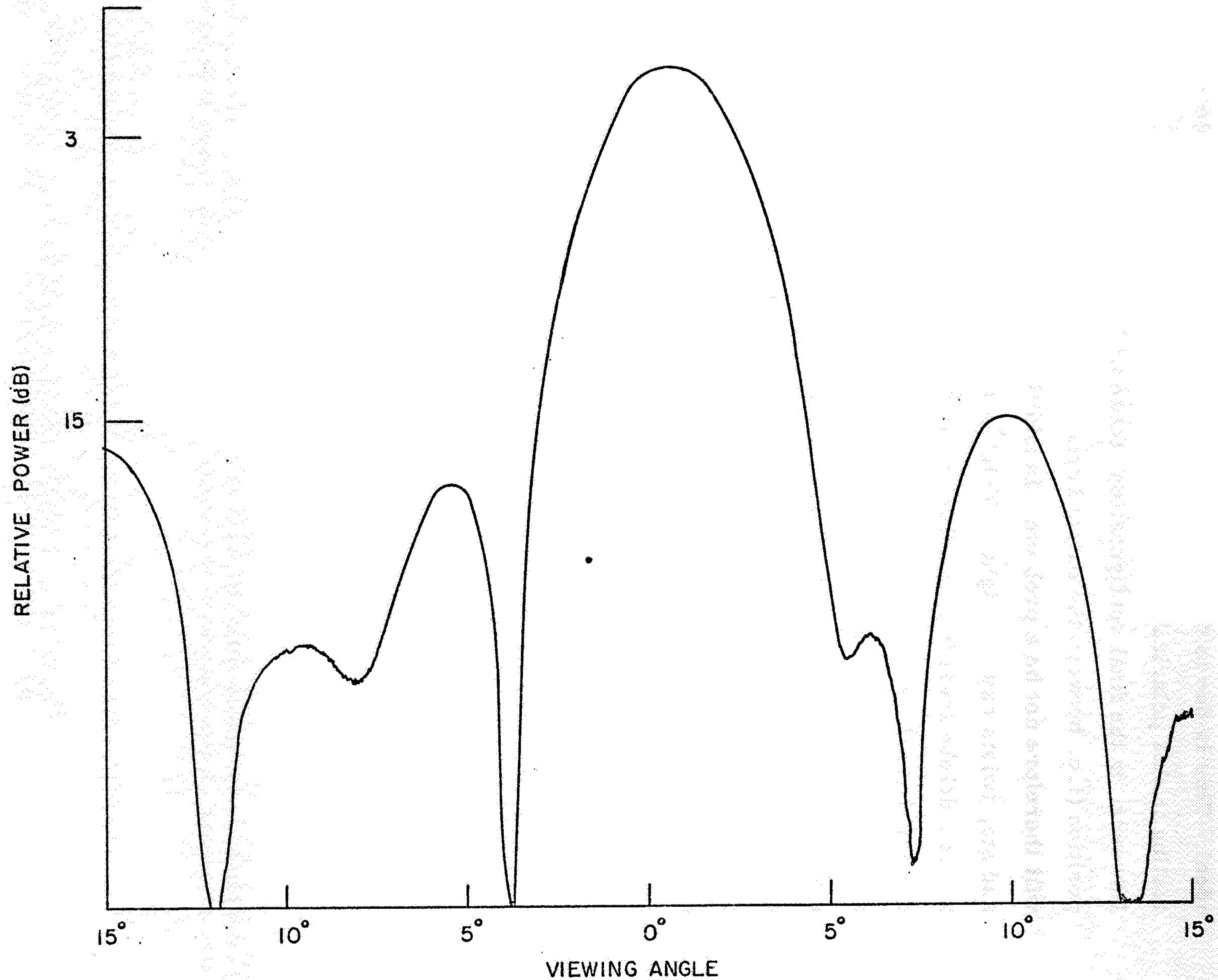

FIG. 50-LINEAR ARRAY H-PLANE POLARIZATION 35.4 GHz BORESIGHT

In the H-plane polarization patterns it is immediately evident that the center frequency patterns are not as close to the ideal. This is due to the fact that the step twists used to enable testing in this polarization also affect the illumination function. However, in the final configuration, these will not be used in this position (i. e. between the output horns and the phase shifters) and will therefore not be a problem. In addition to the phase shifters and step twists causing higher sidelobes there is an inherent increase in the sidelobe level in this polarization due to the fact that the individual horn does not have a uniform amplitude across its aperture in this polarization. Rather, it approaches a cosine function and must go to zero at the edges of each horn to satisfy the boundary conditions. This only increases the sidelobe level in the theoretical patterns by approximately 0.3 dB. In general the patterns are very close to what was expected using the digital phase shifters and when these reach an advanced state of development it is expected that the system will perform very well across the band from 34.5 to 35.5 GHz.

8. CONCLUSIONS AND RECOMMENDATIONS

The substance of the accomplishments of this program can be stated very succinctly. It was determined that an array antenna could satisfy the demands of a satellite communications system. Furthermore, it is especially appealing at millimeter wavelengths where the size and weight reduction possible make it easily adaptable to space applications. It was also resolved that a small number of higher gain elements which completely fill the aperture of the array were adequate to achieve the gain, beamwidth, sidelobe levels, bandwidth and steering capabilities for the antenna. The recommended methods of distributing the millimeter wave energy to these apertures from the transmitter were reduced to two - a series waveguide coupler feed, and a quasi-optical feed. Although the former was demonstrated to be in excellent agreement with predicted performance at 35 GHz it is felt that the higher losses in waveguide with increasing frequency would limit its use at 94 GHz. On the other hand, the reported measurements of the quasi-optical feed, although this was not completely developed, indicate that it holds more promise at higher frequencies where it would be less frequency sensitive and have less loss than the former system. For these reasons it is strongly recommended that development be continued.

Since from synchronous orbit the earth subtends an 8 $1/2^{\circ}$ cone, allowances for some variations in platform orientation led to an antenna system designed with the capability of steering $\pm 12^{\circ}$. The beamwidth, which for the series coupler system, was measured to be 3.8° implies a spot size on the earth of approximately 1500 miles at the nadir angle. Using the formulas developed on the program it is possible to examine conveniently the trade-offs in the parameters of the aperture for an antenna system with different requirements of steering, beamwidth, etc.

For the present application, the antenna system requires one phase shifter per aperture element. The most desirable choice for these phase shifters, subject to continuing development, is a latching ferrite type. The control circuits for steering can then be binary logic circuits which are ideally suited for microcircuitry and help minimize weight and volume problems.

The weights of the respective systems in their present and projected form are given in Appendix I, Table I.

9. NEW TECHNOLOGY

During the period covered by this report, there have been no inventions, discoveries, or innovations which may be considered under the New Technology clause of the contract.

APPENDIX I.

The weight of both systems are given below and are based on the minimum wall thickness compatible with strength. It is also assumed that, where possible all components will be fabricated using electroforming techniques. The present weight of the phase shifters was used in the projected weight of the system since it was not known how much if any reduction could be accomplished. The weight of 49 phase shifters is about 6.4 lbs.

TABLE II

	Present Wt.	Projected Wt. (min)
Optically Fed System		
8 sectoral horns	5.92 lbs.	.38 lbs.
49 collecting horns	3.46 lbs.	.17 lbs.
49 output horns	3.46 lbs.	1.73 lbs.
7 non uniform	.61 lbs.	.02 lbs.
98 flanges	1.48 lbs.	1.48 lbs.
49 waveguide bends	-	.92 lbs.
49 phase shifters	6.40 lbs.	6.40 lbs.
	21.33 lbs.	11.10 lbs.
Series Coupler System		
8 series feeds	5.33 lbs.	2.09 lbs.
49 output horns	3.46 lbs.	1.73 lbs.
49 flanges	.74 lbs.	.74 lbs.
49 phase shifters	6.4 lbs.	6.4 lbs.
7 step twists	.21 lbs.	.21 lbs.
	16.14 lbs.	11.17 lbs.

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